Transmission in multi-channel communication system using selective channel power control

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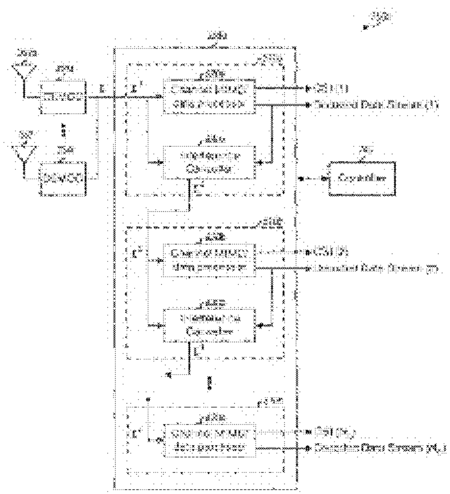
Abstract of corresponding document: **WO 02093779 (A2)**

Techniques to process data for transmission over a set of transmission channels selected from among all available transmission channels. In an aspect, the data processing includes coding data basd on a common coding and modulation scheme to provide modulation symbols and pre-weighting the modulation symbols for each selected channel based on the channel's characteristics. The pre-weighting may be achieved by "inverting" the selected channels so that the received SNRs are approximately similar for all selected channels. With selective channel inversion, only channels having SNRs at or above a particular threshold are selected, "bad" channels are not used, and the total available transmit power is distributed across only "good" channels.; Inproved performance is achieved due to the combined benefits of using only the Ns best channels and matching the received SNR of each selected channel to the SNR required by the selected coding and modulation scheme.



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Radio transmission apparatus and radio communication method

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Inventor(s):	KIMIHIKO ISHIKAWA [JP] +
Applicant(s):	MATSUSHITA ELECTRIC IND CO LTD [JP] +
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Application number:	CN20028004161 20021030
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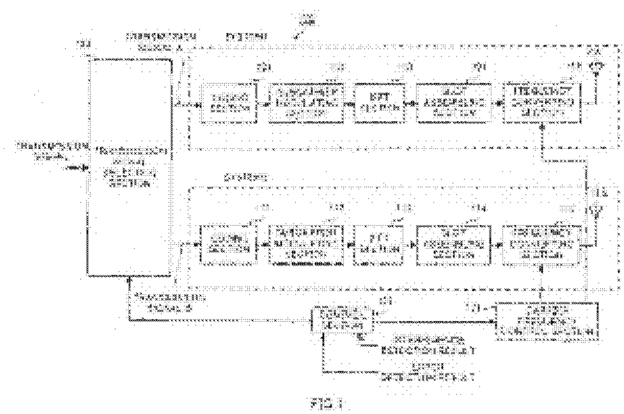
Translate this text

There is provided a radio transmitting apparatus capable of improving spectrum efficiency and a transmission rate while maintaining communication quality. The present apparatus adaptively selects space multiplex where different information (transmission signal A NOTEQUAL transmission signal B) is transmitted from a plurality of antennas with the same frequency, frequency multiplex where different information (transmission signal A NOTEQUAL transmission signal B) is transmitted from the plurality of antennas with different frequencies, space diversity where the same information is transmitted from the plurality of antennas with the same frequency, and frequency diversity where the same information (transmission signal A = transmission signal B) is transmitted from the plurality of antennas with different frequencies according to circumstances of a propagation path.



CN1489836A Radio transmission apparatus and radio communication method

Bibliography

DWPI Title
Radio transmitter switches between transmission of different data with identical or different frequency and transmission of identical data with identical or different frequency, based on transmission path condition

English Title
Radio transmission apparatus and radio communication method

Assignee/Applicant
Standardized: **MATSUSHITA ELECTRIC IND CO LTD**

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KIMIHIKO ISHIKAWA

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[51] Int. Cl⁷
H04B 7/06

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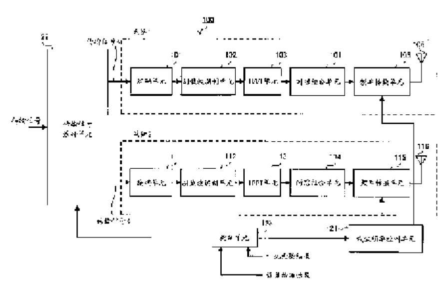
权利要求书 1 页 说明书 7 页 附图 6 页

[54] 发明名称 无线发射装置和无线通信方法

[57] 摘要

一种无线发射装置，能够进一步改善频率使用效率和传输速率，同时保持通信质量。在该装置中，根据传输路径条件来自适应地进行选择：用来自多个天线(106、116)的相同频率来传输不同的信息(传输信号 A≠传输信号 B)(空间复用)；用来自多个天线(106、116)的不同频率来传输不同的信息(传输信号 A≠传输信号 B)(频率复用)；用来自多个天线(106、116)的相同频率来传输相同的信息(传输信号 A = 传输信号 B)(空间分集)；用来自多个天线(106、116)的不同频率来传输相同的信息(传输信号 A = 传输信号 B)(频率分集)。



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1. 一种无线发射装置，包括:

发射装置，使用多个天线来发射相同或不同的信息;

测量装置，用于测量传播路径的情况；以及

控制装置，根据由所述测量装置所测量的传播路径的情况，来控制多个天线的每个发射频率和从多个天线发射的信息数量。

2. 如权利要求1所述的无线发射装置，还包括用于检测未分配给其他用户的频率的检测装置，其中所述控制装置根据由所述检测装置所检测的频率来设置多个天线的每个发射频率。

3. 如权利要求1所述的无线发射装置，其中，当由所述测量装置测量的传播路径的情况良好时，所述控制装置将多个天线的每个发射频率设置为相同的频率，以便从多个天线发射不同的信息。

4. 如权利要求1所述的无线发射装置，其中，当由所述测量装置测量的传播路径的情况差时，所述控制装置将多个天线的每个发射频率设置为不同的频率，以便从多个天线发射不同的信息。

5. 如权利要求1所述的无线发射装置，其中，当由所述测量装置测量的传播路径的情况差时，所述控制装置将多个天线的每个发射频率设置为相同的频率，以便从多个天线发射相同的信息。

6. 如权利要求1所述的无线发射装置，其中，当由所述检测装置测量的传播路径的情况差时，所述控制装置将多个天线的每个发射频率设置为不同的频率，以便从多个天线发射相同的信息。

7. 一种无线基站装置，包括根据权利要求1的无线发射装置。

8. 一种无线终端装置，包括根据权利要求1的无线发射装置。

9. 一种无线通信方法，包括步骤:

使用多个天线来发射相同或不同的信息;

测量传播路径的情况；以及

根据由所述测量步骤测量的传播路径的情况，来控制多个天线的每个发射频率和从多个天线发射的信息数量。

无线发射装置和无线通信方法

技术领域

本发明涉及一种在数字无线通信系统中使用的无线发射装置和无线通信方法。

背景技术

目前正在进行针对世界范围统一标准的宽带无线接入系统的准备。而且，对于下一代，期望建立充分使用接近亚毫米波带的丰富的频率资源的移动宽带无线接入系统。

作为目前的宽带无线接入系统，在所使用的一种系统中调制方法是在世界范围统一标准下使用 5GHz 的频带的正交频分复用（OFDM），并且根据传播路径的情况来自适应地控制对应于每个副载波的调制多值数量。根据这种方法，在好的传播路径情况下，能够获得一个大的调制多值数量。为此，能够获得使用例如 20MHz 的频带中的 64－值 QAM 的 54Mbps 的传输速率。

最近几年，为了改善频率的有效使用，已经考虑了 SDM（空分复用）方法的应用，在 SDM 方法中，使用多个天线用相同的频率来执行空分复用（“宽带移动通信系统的 PDM-COFDM 方案”，Sugiyama，Umehira，电子、信息和通信工程学会，通信协会，2001，SB-3-7）。在这种类型的方法中，调制与常规情况下的调制相同，但是用相同的频率从多个天线发射不同的信息来执行空间复用。为此，例如，在使用两个天线的情况中，传输容量加倍而不会增加将使用的频带，因此也加倍了传输速率。

然而，在上述的常规方法中，存在这样一种情况：根据来自另一个小区的干扰和传播路径的情况，原则上接收方不能执行受到空间复用的传输信号的分离和再现。为此，不能一直增加通信容量，并且存在这样一种可能性，即，将发生不能满足预期的传输速率的需要的这样一种情况。另外，有这样一种可能，根据这种情况将发生通信不可能状态。

发明内容

本发明的一个目的是提供一种无线发射装置和无线通信方法，其能够改

善频频谱效率和传输速率，同时保持通信质量。

本发明的实质是：当使用多个天线来发射相同或不同的信息时，根据传播路径的情况来控制多个天线的每个发射频率和从多个天线发射的信息数量，即，自适应地选择使用相同频率从多个天线发射不同信息的空间复用、使用不同频率从多个天线发射不同信息的频率复用、使用相同频率从多个天线发射的相同信息的空间分集，以及根据传播路径的情况使用不同频率从多个天线发射的相同信息的频率分集。

附图说明

图1示出了根据本发明一个实施例的无线发射装置的结构方框图；

图2示出了执行与图1所示的无线发射装置无线通信的无线接收装置的结构方框图；

图3示出了能够由图1所示的无线发射装置执行的无线通信系统的空间复用的图；

图4示出了能够由图1所示的无线发射装置执行的无线通信系统的频率复用的图；

图5示出了能够由图1所示的无线发射装置执行的无线通信系统的空间分集的图；和

图6示出了能够由图1所示的无线发射装置执行的无线通信系统的频率分集的图。

具体实施方式

将参考附图具体解释本发明的一个实施例。

图1示出了根据本发明一个实施例的无线发射装置的结构方框图，以及图2示出了执行与图1所示的无线发射装置无线通信的无线接收装置的结构方框图。图1示出的无线发射装置和图2示出的无线接收装置能够被安装在相同的无线通信装置中。

这种无线通信装置是一种OFDM无线通信装置，图1中示出的无线发射装置是对于OFDM信号的发射器，图2中示出的无线接收装置200是对于OFDM信号的接收器。由于通过多载波转换和保护间隔插入在高速数字信号发射中能够减小多路延迟扩展的影响，所以需要注意作为下一代的移动宽带

无线接入的 OFDM。这里，OFDM 信号是通过对多个正交副载波信号进行多路复用而获得的信号。

根据本实施例，当使用 OFDM 中的多个天线来发射相同或不同的信息时，根据传播路径的情况来控制多个天线中的每个的发射频率以及从多个天线发射的信息数量，从而改善频率使用效率和传输速率，同时保持了通信质量。下文将解释作为一个示例的天线数量是二的一种情况。

图 1 所示的无线发射装置(发射器)100 包括发射传输信号 A 的系统 1 以及发射传输信号 B 的系统 2。系统 1 包括编码单元 101、副载波调制单元 102、逆快速傅里叶变换(IFFT)单元 103、时隙组合单元 104、频率转换单元 105、天线 106。系统 2 包括编码单元 111、副载波调制单元 112、逆快速傅里叶变换(IFFT)单元 113、时隙组合单元 114、频率转换单元 115、天线 116。而且，发射器 100 包括载波频率控制单元 121、传输信号选择单元 122、和用于整体的控制单元 123。

同时，图 2 所示的无线接收装置(接收器)200 包括：系统 1，接收来自发射器 100 的传输信号 A，以便获得所接收的信号 A；和系统 2，接收来自发射器 100 的传输信号 B，以便获得所接收的信号 B。然而，当传输信号 A 和传输信号 B 具有相同频率时，在每个系统接收传输信号 A 和传输信号 B。系统 1 包括天线 201、频率转换单元 202、逆快速傅里叶变换(FFT)单元 203、副载波解调单元 204、以及解码单元 205。系统 2 包括天线 211、频率转换单元 212、逆快速傅里叶变换(FFT)单元 213、副载波解调单元 214、以及解码单元 215。而且，接收器 200 包括载波频率控制单元 221、符号同步定时单元 222、干扰补偿单元 223、干扰检测单元 224、以及误差检测单元 225。

此外，当无线发射装置 100 和无线发射装置 200 被安装在相同的无线通信装置上时，发射器 100 的天线 106 和 116 以及接收器 200 的天线 201 和 211 可以是发射和接收共享类型。

下一步将解释上述结构的发射器 100 和接收器 200 的各自操作。

首先，下面将描述发射器 100 的操作。

例如，系统 1 的传输信号 A 通常是由编码单元 101 编码的。对于每个副载波，由副载波调制单元 102 来调制所编码的信号，并在其后将该结果输出到 IFFT 单元 103。IFFT 单元 103 对副载波调制单元 103 的一个输出信号进行逆快速傅里叶变换(IFFT)，以便产生一个 OFDM 信号。通过时隙组合单元 104，

将保护间隔和前同步信号插入已产生的OFDM信号，之后将结果输出到频率转换单元105。频率转换单元105将时隙组合单元104的一个输出信号上变频为一个无线频率(传输频率)，该频率由载波频率控制单元121独立控制。上变频的传输信号通过天线106发射。而且，对系统1的传输信号A进行与系统2的传输信号B同样的处理，频率转换单元115将时隙组合单元114的一个输出信号上变频为一个无线频率(传输频率)，该频率由载波频率控制单元121控制独立，并且之后从天线116发射结果。这时，通过传输信号选择单元122来选择是否系统1的传输信号A和系统2的传输信号B被设置为相同或不同。通过控制单元123来自适应地控制载波频率控制单元121和传输信号选择单元122。由控制单元123做出的自适应控制的内容将在后面具体描述。

下面将描述接收器200的操作。

通过频率转换单元202使用无线频率(与发射传输信号A的天线106的传输频率相同的频率)来下变频经系统1的天线201接收的OFDM信号，该无线频率由载波频率控制单元221独立控制，并且之后经保护间隔移除单元(未示出)将结果输出到FFT单元203。使用从符号同步定时单元222输出的定时信号，FFT单元203对受到保护间隔移除的OFDM信号进行快速傅里叶变换(FFT)。对经系统2的天线211接收的OFDM进行相同的处理。通过频率转换单元212使用一个无线频率(与发射传输信号B的天线116的传输频率相同的频率)来下变频OFDM信号，该无线频率不受载波频率控制单元221的控制，并且之后从其中移除保护间隔，并进行FFT处理。干扰补偿单元223估计天线201和211之间的传递函数，以便分离受到空间复用的信号。在系统1和系统2中，对于每个副载波，通过副载波解码单元204和214对所分离的信号进行解码。从而获得所接收的信号A和所接收的信号B。

这时，干扰测量单元224测量每一系统的干扰电平，并检测干扰波的存在和缺少，从而根据多个可用的频率来检测未分配给其它用户的频率。而且，误差检测单元225检测一个作为指示传播路径情况的基准的误码率(例如，BER(位误码率)等)，经安装在相同无线通信装置上的发射器(未示出)和安装在通信另一端的无线通信装置的接收器(未示出)，将由干扰检测单元224执行的干扰检测结果(每个系统的干扰波的存在或不存在)以及由误差检测单元225执行误差检测结果(每个系统的误码率)传输到通信另一端的发射器100的控

制单元 123。

参考图 3 到 6，接下来给出在发射器 100 的上述自适应控制的内容的解释。另外，在下文解释作为示例的使用包括信道 1(CH1)到信道 4(CH4)的四个信道(频带)的情况。而且，在图 3 到 6 中，天线＃1 表示系统 1 的天线 106，天线＃2 表示系统 2 的天线 116。

在这个实施例中，发射器 100 能够使用四个无线通信系统。第一种情况是空间复用情况，即，一种用相同频率(例如图 3)从两个天线 106 和 116 发射不同信息(传输信号 A≠传输信号 B)的情况。第二种情况是频率复用情况，即一种用不同频率(例如图 4)从两个天线 106 和 116 发射不同信息(传输信号 A≠传输信号 B)的情况。第三种情况是空间分集情况，即，一种用相同频率(例如图 5)从两个天线 106 和 116 发射相同信息(传输信号 A＝传输信号 B)的情况。第四种情况是频率分集情况，即，一种用不同频率(例如图 6)从两个天线 106 和 116 发射相同信息(传输信号 A＝传输信号 B)的情况。根据来自安装在相同无线通信装置上的接收器 200 的干扰检测结果(每个系统的干扰波的存在或缺少)和误差检测结果(每个系统的误码率)，控制单元 123 在这四种无线通信方法中进行自适应地切换。

具体地，例如，当误差检测结果良好时，即传播路径的情况良好，如图 3 所示，用相同的频率从两个天线 106 和 116 发射不同的信息(传输信号 A≠传输信号 B)，从而执行空间复用。在图 3 示出的示例中，避免了存在干扰波的 CH1、CH2、CH4，即分配到其它用户的频率(信道)，并且不同的传输信号 A 和 B 被复用，并使用空闲的相同信道(CH3)分别从系统 1 的天线 106 和系统 2 的天线 116 被发射。另外，这时，接收器 200 使用由发射器 100 使用的频率(图 3 的示例中的 CH3 的频率)来执行接收操作。

根据该方法，当传播路径的情况良好时，执行空间复用，因此能够将频谱效率和传输速率进一步改善到最大程度，而不会增加将被使用的频率，即同时保持要使用的频率。此外，根据多个(这种情况中是四个)可用的频率(CH1 到 CH4)来检测不带频率波的频率，并且根据所检测的频率来设置多个天线(这种情况中是二个)的每个的传输频率。为此，频谱效率和传输速率能够进一步改善，而没有来自其它用户的干扰的影响，即同时保持通信质量。

而且，例如，当误差检测结果差时，即传播路径的情况差，如图 4 所示，用不同的频率从两个天线 106 和 116 发射不同的信息(传输信号 A≠传输信号

B)，并且从而执行频率复用。在图 4 示出的示例中，避免了存在干扰波的 CH1、CH4，即，分配到其它用户的频率(信道)，使用空闲信道 CH2 和 CH3 中的一个信道(CH2)从系统 1 的天线 106 发射传输信号 A，并且使用与系统 1 不同的另一信道(CH3)从系统 2 的天线 116 来发射不同于系统 1 的传输信号 B。另外，这时，接收器 200 使用由发射器 100 使用的每个系统的频率(图 4 示例中系统 1 使用 CH2 的频率和系统 2 使用 CH3 的频率)来执行接收操作。

根据该方法，当传播路径的情况良好时，执行频率复用，因此能够进一步改善谱频率和传输速率，由于每个系统的空间复用而没有通信质量恶化的影响，即，同时保持通信质量。此外，根据多个(这种情况中是四个)可用的频率(CH1 到 CH4)来检测没有频率波的频率，并且根据所检测的频率来设置多个天线(这种情况中是二个)的每个的传输频率。为此，频谱效率和传输速率能够进一步改善，而没有来自其它用户的干扰的影响，即同时保持通信值。

而且，例如，当误差检测结果非常差时，即传播路径的情况如此差以致不能从多个天线传输不同的信息，使用与图 5 所示相同的频率从两个天线 106 和 116 选择性地传输相同的信息(传输 A = 传输 B)，从而执行空间分集。或者，用图 6 所示不同的频率从两个天线 106 和 116 传输相同的信息(传输 A = 传输 B)，从而执行频率分集。在图 5 所示的示例中，避免了存在干扰波的 CH1、CH2、CH4，即分配到其它用户的频率(信道)，而且相同的传输信号(传输 A = 传输 B)是使用空闲相同信道(CH3)从系统 1 的天线 106 和系统 2 的天线 116 传输的空间分集。在图 6 所示的示例中，避免了存在干扰波的 CH1 和 CH4，即分配到其它用户的频率(信道)，并且传输信号是使用两个空闲信道 CH2 和 CH3 中的一个信道(CH2)从系统 1 的天线 106 传输的，并且与系统 1 相同的传输信号(传输信号 A = 传输信号 B)是使用与系统 1 不同的另一信道(CH3)来传输的。另外，这时，在前面的情况中，接收器 200 使用由发射器 100 使用的每个系统频率(图 5 示例中的 CH3 的频率)来执行接收操作。在后面的情况中，接收器 200 使用由发射器 100 使用的频率(图 6 示例中系统 1 使用 CH2 的频率和系统 2 使用 CH3 的频率)来执行接收操作。

根据该方法，当传播路径的情况非常差时，即，需要以传输速率的改善为代价来确保通信质量，执行空间分集或频率分集。为此，即使传播路径的情况如此差以致不能从多个天线传输不同的信息，也能够通过分集来维持通信质量。

因此，根据本实施例，当使用多个天线106和116来传输相同或不同的信息时，根据传播路径的情况来控制多个天线106和116的每个传输频率以及从多个天线106和116发射的信息数量，例如，根据传播路径的情况来自适应地选择空间复用、频率复用、空间分集和频率分集，并且这样能够进一步改善频谱效率和传输速率，同时保持通信质量。换句话说，能够同时获得通信质量的保持以及谱频率的进一步改善。

此外，根据本实施例，发射器100的自适应控制基于这种构思：即使存在空闲频率，首先使用相同的频率(为了其它用户以后能够容易地访问)，并且当通信质量不能得到保证时，使用不同的频率(然而，有必要检测不存在干扰波)。然而，自适应控制的控制概念并不限于此。

例如，能够采用这种构思：当没有干扰波存在时，不管传播路径的情况是否好或坏，用不同的频率来执行传输。具体地，例如，可以考虑下列使用。即，首先检测干扰波的存在与否，并且当干扰波不存在时，使用不同的频率，当操作期间检测到干扰波时，使用相同的频率。然后，当检测到干扰波在后来消失时，再次使用不同的频率。在这种情况中，由于不管传播路径的情况而使用没有干扰波的频率，将多个天线的每个传输频率设置为不同的频率，所以能够随便使用未分配的，即空闲的频率，并且能够减小来自其它用户的干扰的影响。

另外，本实施例解释了作为一个示例的OFDM无线通信装置。然而，本发明并不限于OFDM系统的应用。例如，本发明也能够应用于CDMA(码分多址)无线通信装置。

而且，本发明的无线发射装置能够安装在无线通信装置上，例如，移动通信系统中的无线基站装置和无线终端装置。

如上所解释的，根据本发明，能够进一步改善频谱效率和传输速率，同时保持通信质量。

本申请基于2001年10月31日提交的日本专利申请号2001-334392，在此全文引用，以供参考。

工业实用性

本发明可应用于无线装置，例如移动通信系统中的移动站装置和无线基站装置。

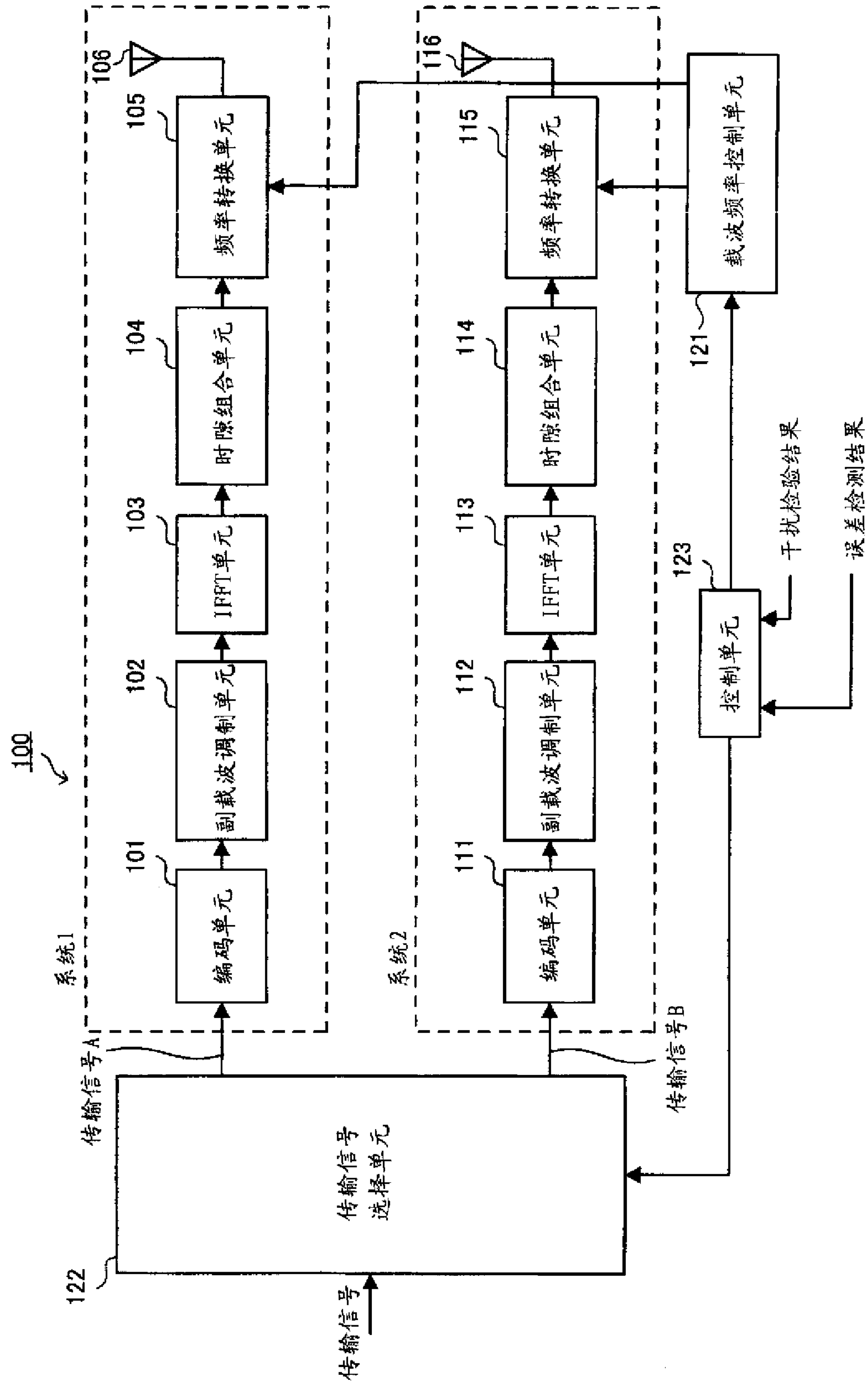

100
系统1
系统2
传输信号
传输信号选择单元
122
传输信号A
传输信号B
编码单元
101
111
副载波调制单元
102
112
IFFT单元
103
113
时隙组合单元
104
114
频率转换单元
105
115
106
116
载波频率控制单元
121
控制单元
123
干扰检验结果
误差检测结果

图 1

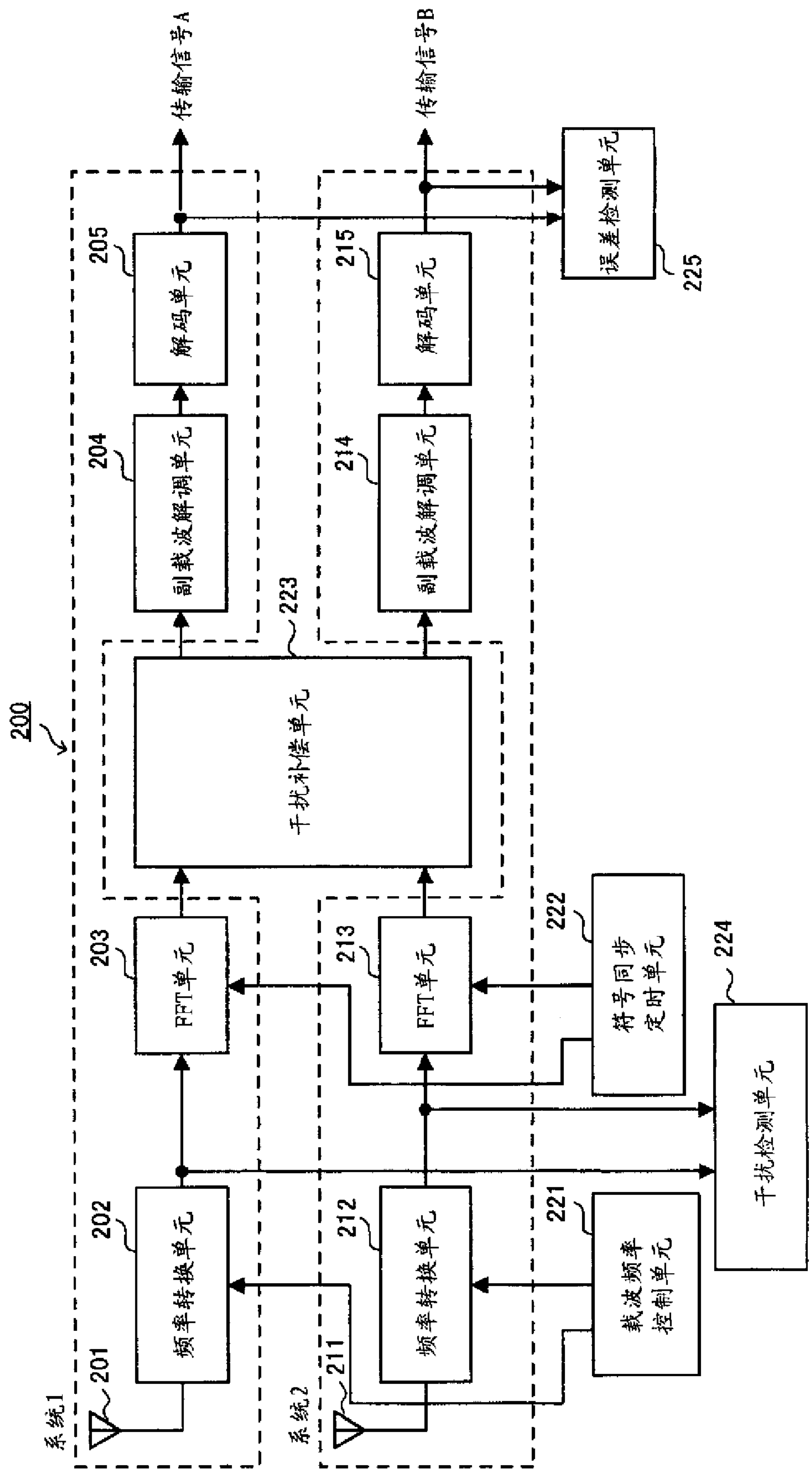

200
系统1
201
202
频率转换单元
203
FFT单元
223
干扰补偿单元
204
副载波解调单元
205
解码单元
传输信号A
系统2
211
212
频率转换单元
213
FFT单元
214
副载波解调单元
215
解码单元
传输信号B
221
载波频率控制单元
222
符号同步定时单元
224
干扰检测单元
225
误差检测单元

图 2

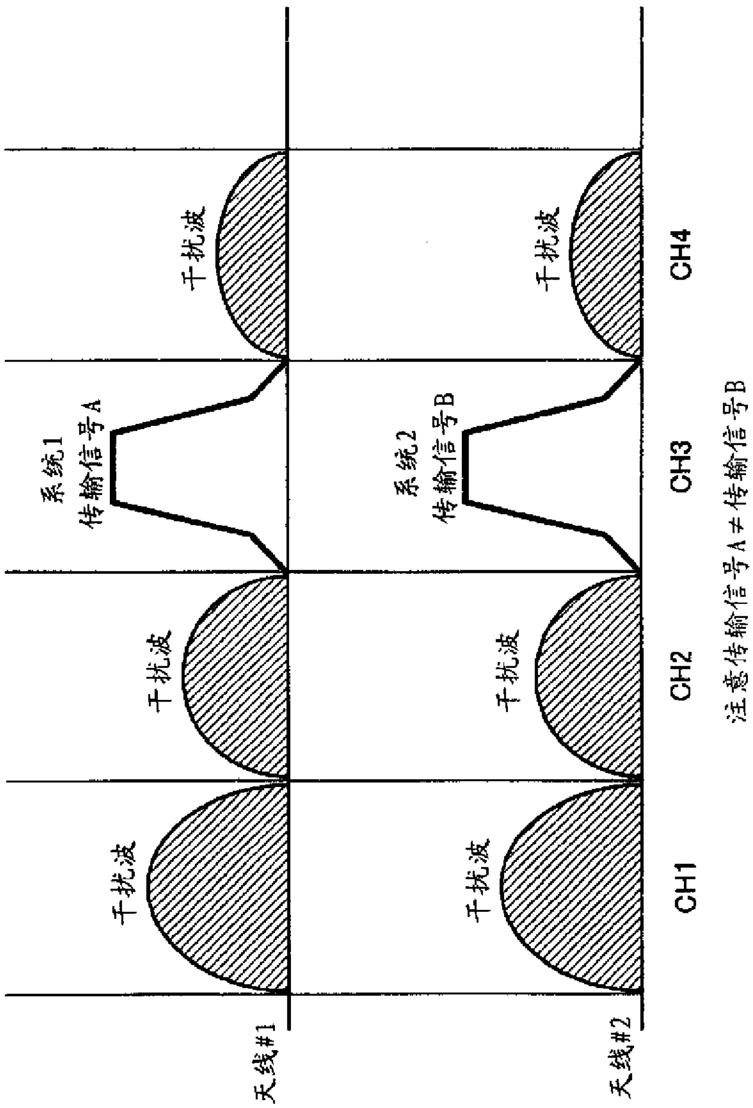
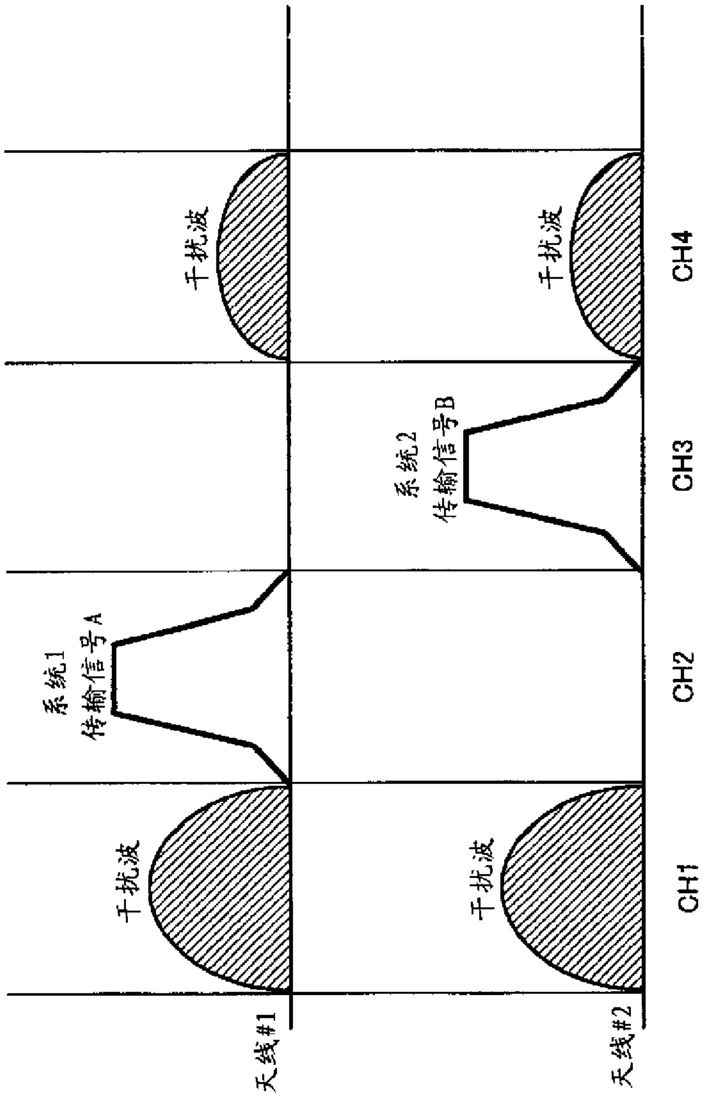

天线#1
天线#2
干扰波
干扰波
干扰波
干扰波
干扰波
干扰波
系统1
传输信号A
系统2
传输信号B
CH1
CH2
CH3
CH4
注意传输信号A≠传输信号B

图 3



注意传输信号A≠传输信号B

图 4

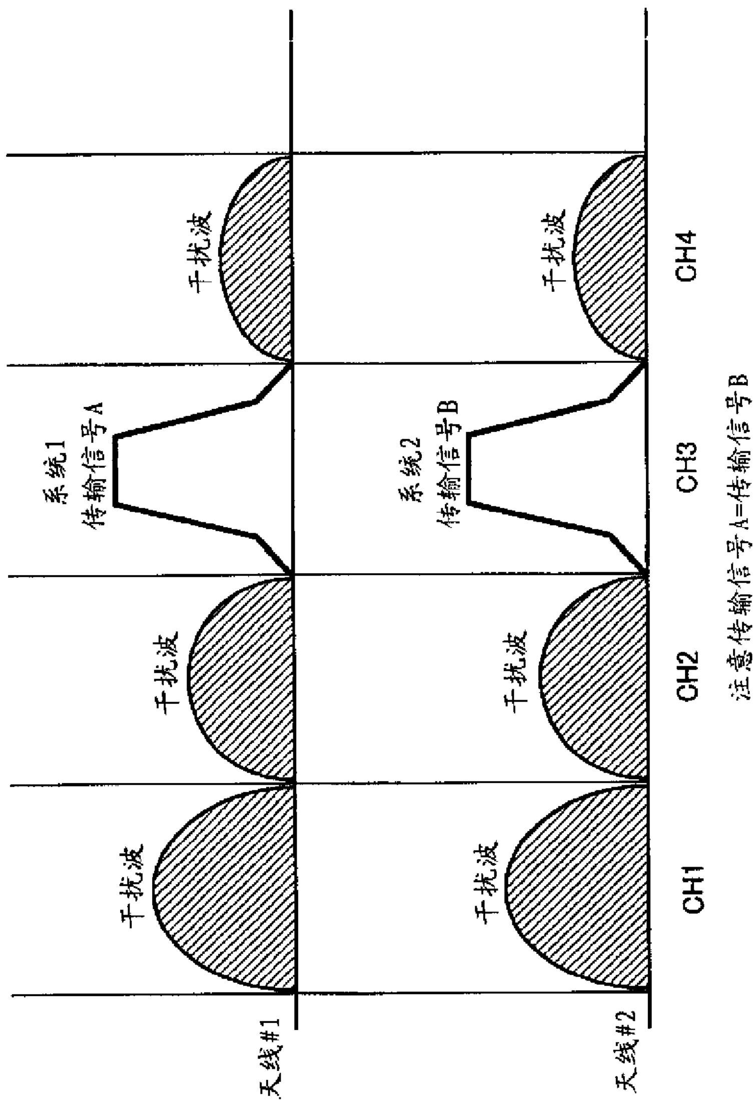

天线#1
天线#2
干扰波
干扰波
干扰波
干扰波
干扰波
干扰波
系统1
传输信号A
系统2
传输信号B
CH1
CH2
CH3
CH4
注意传输信号A=传输信号B

图 5

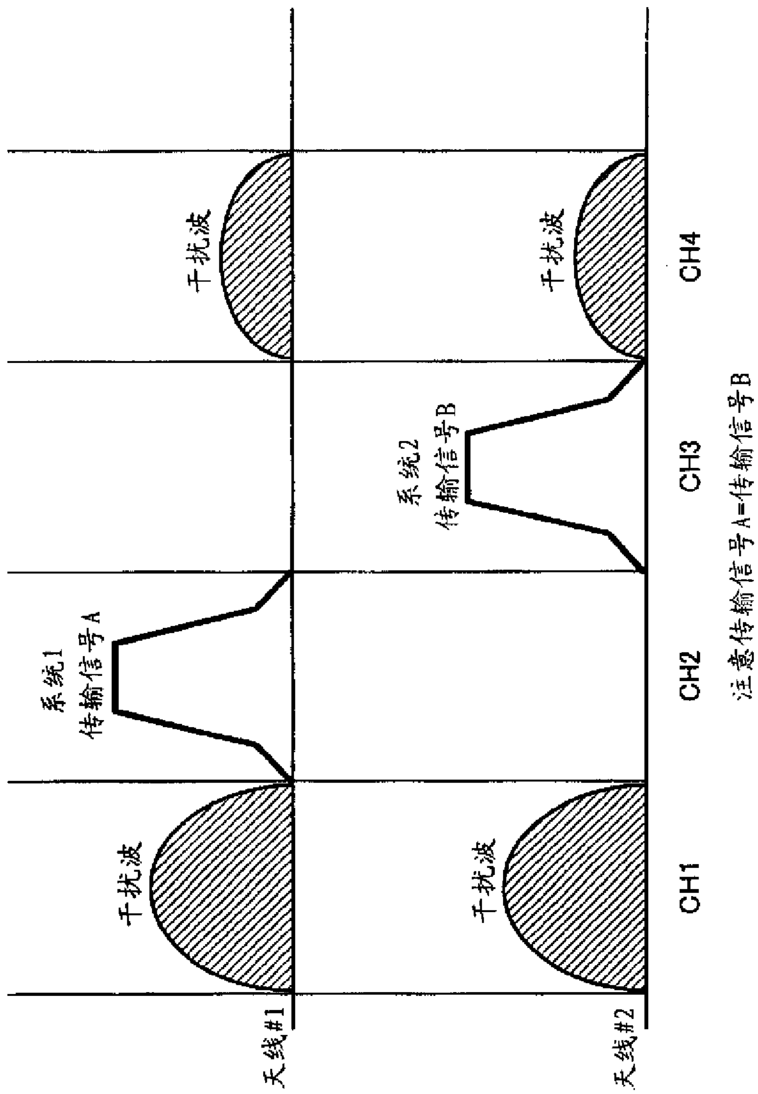

天线#1
天线#2
干扰波
干扰波
干扰波
干扰波
系统1
传输信号A
系统2
传输信号B
CH1
CH2
CH3
CH4
注意传输信号A=传输信号B

图 6

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European Patent Office
Office européen des brevets

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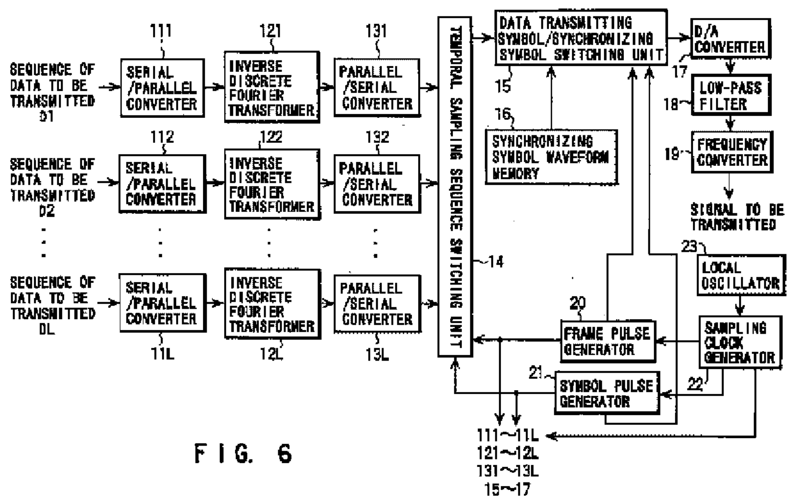
(54) MCM system, with a plurality of data streams, each with its own transmission parameters

(57) Data sequences (D1 through DL) correspond to L different parameter sets (effective symbol length, guard interval length, the number of carrier waves) and transformed into parallel data by respective serial/parallel converters (111 through 11L), which are allocated to respective carrier waves for OFDM and subjected to inverse discrete Fourier transform by inverse discrete Fourier transformers (121 through 12L) to produce sampled values for the transmission waveform in the time domain. The sampled values are transformed into serial sequences of sampled values by parallel/serial converters (131 through 13L) and then into a single temporal sampling sequence by a temporal sampling sequence switching unit (14). A frame synchronizing symbol is added to the temporal sampling sequence and then transformed into an analog base band OFDM signal before it is converted up to a transmission signal. The frequency bandwidth of the OFDM signal is made smaller than a predetermined value defined by the bandwidth of the available transmission channel. As a result, an OFDM signal that can be received well regardless of the mode of reception can be transmitted.



F I G. 6

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Office européen des brevets

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(54) Multicarrier modulation system, with variable symbol rates

(57) An OFDM system uses a normal mode which has a symbol length T, a guard time T_G and a set of N sub-carriers, which are orthogonal over the time T, and one or more fallback modes which have symbol lengths KT and guard times KT_G where K is an integer greater than unity. The same set of N sub-carriers is used for the fallback modes as for the normal mode. Since the same set of sub-carriers is used, the overall bandwidth is substantially constant, so alias filtering does not need to be adaptive. The Fourier transform operations are the same as for the normal mode. Thus fallback modes are provided with little hardware cost. In the fallback modes the increased guard time provides better delay spread tolerance and the increased symbol length provides improved signal to noise performance, and thus increased range, at the cost of reduced data rate.

EP 0 929 172 A1

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Office européen des brevets

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(54) Web-based enterprise management with multiple repository capability

(57) A transport neutral technique allows an object manager (20) to communicate with a CIM repository (130) using any of a variety of protocols (158, 160). The object manager software (20) is independent of the transport mechanism used and need not be changed if the transport mechanism changes. A computer system (110) to be managed includes a CIM object manager (20) and any number of provider APIs (122) that provide resource information about the computer system. A CIM repository (130) stores classes and instances used by the object manager. A remote application computer (150) runs a software management application (32) that communicates with the object manager of the computer system using a local client API (156). A Repository API of the object manager includes an interface definition (300) defining all methods called by the object manager. Also included is a protocol-specific class that implements the interface definition; there is a protocol-specific class for each protocol desired to be supported. Each class implements methods using a specific protocol. A factory class is executed when the object manager invokes a method call passing in a desired protocol parameter. The factory class creates a protocol-specific object of one of the protocol-specific classes depending on the protocol parameter. The object is returned to the object manager which executes one of its protocol-specific methods thus allowing communication to a repository using a protocol independent of the object manager.

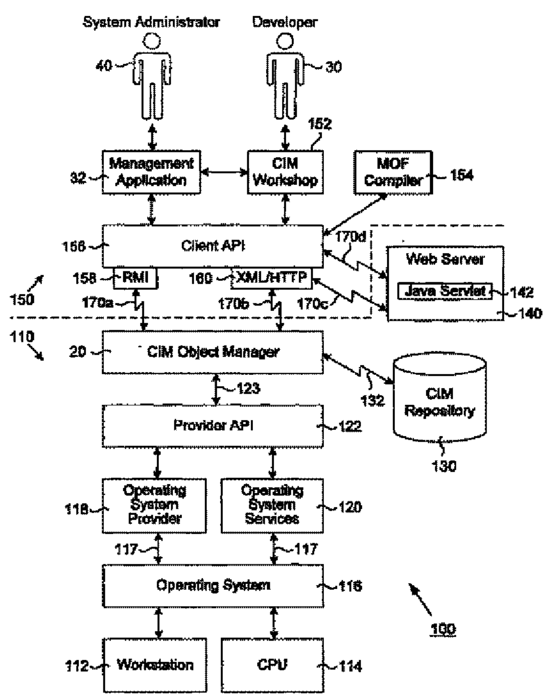

FIG. 2A

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2.16.7 (HRS)/3.6

Description

FIELD OF THE INVENTION

[0001] The present invention relates generally to managing the resources of a computer system. More specifically, the present invention relates to a technique for communicating database operations from a Common Information Model (CIM) object manager to multiple remote CIM repositories.

BACKGROUND OF THE INVENTION

[0002] Recently, computers and their associated peripheral equipment (a computer system) have become increasing more complex. As such, it has become progressively more and more complicated for a user or system administrator to manage the resources of such a computer system. With a variety of peripheral devices and software applications available for use, and their ever-changing nature, the job of a system administrator has become more difficult. Computer system resources such as attached devices, network connections, software applications, etc., must all be managed to ensure an efficiently working system for the user. Within a large corporation having large numbers of such computer systems spread around the world, the task of managing the resources of each computer system can be daunting.

[0003] Recently, the industry has responded to such a need by introducing Web-Based Enterprises Management (WBEM) which is both an initiative and a technology. As an initiative, WBEM includes a standard for managing systems, networks, users, and applications by using Internet technology. As a technology, WBEM provides a way for management applications to share management data independently of vendor, protocol, operating system, or management standard. By developing management applications according to WBEM principles, vendors can develop products that work together easily at a lower cost of development.

[0004] One known standard for implementation of WBEM is the Common Information Model (CIM). CIM is an approach to managing systems and networks. CIM provides a common conceptual framework to classify and define the parts of a network environment and depict how they integrate. The model captures notions that are applicable to all areas of management, independent of technology implementation.

[0005] WBEM software includes tools and technology that software developers can use to create CIM-compliant software applications that manage the environment of a computer system. Developers can also use this software to write "providers," programs that supply data and events for managed objects that are specific to their domain.

[0006] There can be drawbacks, however, associated with various implementations of WBEM software. For example, it may be necessary for the object manager of a computer system to be able to access different types of databases, whether local or remote. Not all implementations, however, are well-suited for this type of access.

[0007] FIG. 1 illustrates a prior art computer system 10 that has resources to be managed. Resources include disk usage, CPU utilization, running applications, etc. Not shown for simplicity are hardware components of the computer system or other software applications. A CIM object manager 20 is responsible for handling all communication between management applications, a CIM repository 26 and managed objects. In this example, CIM repository 26 is a local drive that communicates via a local connection 28 to object manager 20. A provider application programming interface (API) 22 provides an interface to any needed system information 24 to be provided to object manager 20. Object manager 20 may also access CIM repository 26 over local connection 28 to quickly retrieve data objects that have been previously stored. In this fashion, object manager 20 is well-suited for gathering resource information regarding computer system 10.

[0008] In order to efficiently manage the resources of computer system 10, a software developer 30 writes management application software 32 for managing the resources of the computer system. When in operation, the results of management application 32 may be used by a system administrator 40 to manage the computer system. Client application 32 communicates via a Client API 34 to retrieve resource information from computer system 10. Client API 34 uses any suitable local or remote network connection 36 to access object manager 20.

[0009] Prior art implementations of this sort use a single protocol for communication from object manager 20 to CIM repository 26 over local connection 28. Such an implementation is inflexible in that the object manager commands to repository 26 are dependent upon a single protocol. In other words, the commands are not independent of the protocol; should the protocol be modified or if another protocol be used or desired, it will be necessary to rewrite portions of object manager 20 which would be undesirable.

[0010] In addition, having an object manager that is depended upon a particular protocol presents difficulties when repository 26 is remote from object manager 20. In this scenario, it may be desirable to communicate over a network connection using any of a variety of protocols instead of always being required to use a local protocol. In prior art computer system 10 portions of object manager 20 would have to be rewritten for each and every different protocol that is desired to be used.

[0011] Therefore, a technique is desired that would permit an object manager to communicate both locally and remotely with any number of repositories using any of a variety of protocols. It is desired to implement this technique with the least impact upon developers of

object manager software.

SUMMARY OF THE INVENTION

[0012] To achieve the foregoing, and in accordance with the purpose of the present invention, a technique is disclosed that allows an object manager to communicate with any number of repositories using any of a variety of local or remote protocols. Advantageously, the object manager becomes independent of protocol used and need not be changed if the protocol changes.

[0013] In one embodiment, a method is used for communication between a Common Information Model (CIM) object manager and a CIM repository. The method involves first creating a connection between the object manager the CIM repository. Next, a protocol indicator is passed from the object manager to a repository API. The protocol indicator identifies a protocol by which the object manager desires to communicate with the CIM repository. A protocol-specific object is created having methods implemented using the protocol. Finally, the protocol-specific object is returned to the object manager, thus the object manager may communicate with the CIM repository using the protocol desired.

[0014] In another embodiment, a computer system interacts with a CIM repository on a separate computer. The computer system includes an object manager that has program code for interacting with the CIM repository and a protocol indicator. Also included is a repository application programming interface (repository API) that has a factory class arranged to receive the protocol indicator from the object manager and produce a protocol-specific object. Also within the repository API is a first class having methods defined thereon implemented in a first protocol and a second class having methods defined thereon implemented in a second protocol. Thus, the protocol-specific object may be returned to the object manager for use in communicating with the CIM repository using a desired protocol.

BRIEF DESCRIPTION OF THE DRAWINGS

[0015] The invention, together with further advantages thereof, may best be understood by reference to the following description taken in conjunction with the accompanying drawings in which:

FIG. 1 illustrates a prior art computer system that has managed resources.

FIGS. 2A and 2B illustrate a Web-Based Enterprise Management (WBEM) architecture suitable for implementing an embodiment of the invention.

FIGS. 3A and 3B illustrate an example graphical user interface for the CIM workshop of FIG. 2.

FIG. 4 illustrates an interface definition useful for implementing an embodiment of the invention.

FIG. 5 illustrates an implementation definition for the interface of FIG. 4.

FIG. 6 illustrates another implementation definition for the interface of FIG. 4.

FIG. 7 is a JAVA factory class definition.

FIG. 8 is a flowchart illustrating a store or retrieve repository method issued by an object manager to a CIM repository.

FIGS. 9 and 10 illustrate a computer system suitable for implementing embodiments of the present invention.

DETAILED DESCRIPTION OF THE INVENTION

[0016] FIGS. 2A and 2B illustrate a Web-Based Enterprise Management (WBEM) architecture suitable for implementing an embodiment of the invention. Architecture 100 includes computer system 110 having resources to be managed, and application computer 150 where management applications are developed and run. Computer system 110 may be any suitable computer having resources needing to be managed such as CPU load, disk space, installed applications, etc. The hardware layer includes workstation 112 and CPU 114. Workstation 112 may be any suitable computer workstation such as the SPARC workstation from Sun Microsystems, Inc. CPU 114 may be any suitable processor such as an Intel processor. The software layer of computer system 110 includes operating system 116 (other software applications not shown for simplicity) which may be any suitable operating system such as the Solaris operating system from Sun Microsystems, Inc.

[0017] The object provider layer of computer system 110 includes operating system provider 118, operating system services 120 and provider application programming interface (API) 122. The provider layer communicates with operating system 116 via interface 117, for example, by using a JAVA Native Interface (JNI). Object providers act as intermediaries between CIM object manager 20 and one or more managed devices. When object manager 20 receives a request from a management application 32 for data that is not available in CIM repository 130 it forwards the request to a provider. Object providers are installed on the same machine as object manager 20. Object manager 20 then uses provider API 122 to communicate with locally installed providers. Providers are classes that perform various functions in response to a request from object manager 20. For example, providers map information from a managed device to a CIM JAVA class and map

information from a CIM JAVA class to a managed device format.

[0018] Provider API 122 is an API used by various provider programs to communicate information about managed objects to object manager 20. Operating system provider 118 is a collection of JAVA classes or native methods that represent the operating environment of computer system 110. Operating system services 120 also provide logging information from operating system 116 to object manager 20.

[0019] A variety of other providers may also be present. For example, an SNMP provider includes JAVA classes that map CIM data to SNMP data. Also, a CPU-specific provider may be used to transport resource information directly between CPU 114 and provider API 122, thus bypassing operating system 116.

[0020] Providers may be categorized into three types according to the requests they service. An instance type supplies dynamic instances of a given class and supports the retrieval, enumeration, modification and deletion operations. A property type supplies dynamic property values, for example, disk space. A method type supplies methods of one or more classes. A single provider can support both methods and instances. Most providers are "pull" providers which mean they maintain their own data, generating it dynamical if necessary. Pull providers have minimal interaction with object manager 20 and CIM repository 130. The data managed by a pull provider typically changes frequently, requiring the provider to either generate the data dynamically or retrieve it from a local cache whenever an application issues a request. A single provider can act simultaneously as a class instance and method provider by proper registration and implementation of all relevant methods.

[0021] The management layer of computer system 110 includes CIM object manager 20, CIM repository 130 and web server software 140. Object manager 20 may be any suitable WBEM compliant manager. Object manager 20 manages CIM objects that are represented internally as JAVA classes. Client computers running management applications (such as application computer 150) connect to object manager 20 for resource information about computer system 110. When a WBEM client connects to object manager 20 it receives a reference to that object manager. The client can then perform WBEM operations using this reference.

[0022] When management application 32 uses client API 156 to request or update information about a managed object, object manager 20 contacts either the appropriate provider for that object or a suitable persistent storage mechanism such as repository 130. In one embodiment, classes that are handled by a provider have a "provider" qualifier that identifies the provider to contact for the class. When object manager 20 receives a request for a class that has a "provider" qualifier, it routes the request to the specified provider. If no provider is specified it routes the request to repository 130 using JAVA Naming and Directory Interface (JNDI) 132.

[0023] Object manager 20 also performs various start-up functions: starting and registering the RMI server; registering the XML server; setting up a connection to repository 130; and waiting for incoming requests. Object manager 20 also performs other normal operations: performing security checks such as authentication and authorization; performing syntactical and semantic checking of CIM data operations; routing requests to providers or persistent storage; and delivering data from providers or from persistent storage to client management applications.

[0024] CIM repository 130 is a central storage area for CIM class and instance definitions that communicates with object manager 20 via connection 132. Connection 132 may be any suitable local connection within computer system 110, or may be a remote connection. Connection 132 may use any suitable protocol. Further details on communication with repository 130 are provided in FIG. 2B and in FIGS. 4-8.

[0025] Web server software 140 may be any suitable WBEM XML-compliant web server such as the Sun Web Server available from Sun Microsystems, Inc. JAVA servlet 142 converts XML data to the client API format. For example, if management application 132 contains XML data, client API client 156 encodes the data as XML messages and transports the encoded messages to web server 140 that is running JAVA servlet 142. Web server 140 listens for XML messages on a standard port and passes control to servlet 142 when detected. Servlet 142 then decodes the XML messages it receives. Servlet 142 then converts the XML data to the client API format and transmits the information back to client API 156 in RMI format. Alternatively, should object manager 20 support the HTTP format, client API 156 may communicate directly to object manager 20 without the need for web server 140.

[0026] The application layer of WBEM architecture 100 includes management application 32, CIM workshop 152, MOF compiler 154 and client application programming interface (API) 156. In this embodiment of the invention, these elements of the application layer are shown running on application computer 150 (other hardware and software not shown for simplicity). Alternatively, application computer 150 and computer system 110 may be the same computer or the elements of the application layer may reside on a variety of computers and not exclusively on application computer 150.

[0027] A software developer 30 uses any suitable software tool to develop a management application 32 for processing and displaying data from managed objects of computer system 110. Management application 32 uses client API 156 to request information about managed objects from object manager 20. In this fashion, analysis of the resources of computer system 110 can be presented to a system administrator 40 for proper action.

[0028] Client API 156 and provider APIs represent

and manipulate CIM objects. These APIs represent CIM objects as JAVA classes. An object is a computer representation or model of a managed resource of computer system 110 such as a printer, disk drive or CPU. A developer uses the CIM specification to describe managed objects and to retrieve information about managed objects in computer system 110. One advantage of modeling managed resources using CIM is that those objects can be shared across any system that is CIM compliant.

[0029] Management application 32 may be any of a wide variety of software applications written to analyze and manage the resources of computer system 110. By way of example, management application 32 manages system aspects such as disk information (space available, partitions, etc.), CPU load, event processing, date, time, time zone, memory available, ports, etc.

[0030] Application 32 may also manage specific devices of the computer system such as disks, tape drives, modems, other I/O devices, NICs, and network aspects of the system such as TCP/IP, Netbeui, Novell, etc. Further, management application 32 manages the software applications running on computer system 110 by determining what is currently running on the system, what is currently installed, the state of installation, which applications can be terminated, performing application metering, managing application life cycle, process management, user management, etc.

[0031] In one embodiment of the invention, developer 30 uses a CIM workshop 152 written in JAVA for viewing, changing, adding and deleting CIM classes and instances. CIM workshop 152 provides a graphical user interface for the developer. For example, developer 30 may view and select namespaces, may add namespaces, add properties, qualifiers and methods to new classes, view and create instances, and view and modify instance values. Developer 30 may also use CIM workshop 152 to browse a class inheritance tree and change the root of an object tree for a namespace.

[0032] MOF compiler 154 pares files created in the Managed Object Format (MOF), converts files to JAVA class and stores the extracted classes and instances in repository 130. The MOF language is a syntax for defining CIM classes and instances and is described in the CIM specification. Although classes and instances can also be added through client API 156 using JAVA, MOF compiler 154 eliminates the need to write such code. Compiler 154 provides developers and administrators with a simple and fast technique for modifying repository 130.

[0033] In one embodiment, client API 156 is a public API that JAVA applications use to request operations from object manager 20. Client API 156 is used by management application 32 to transfer data to and from object manager 20. Client API 156 includes a variety of classes, instances and methods useful for communicating with object manager 20 using any suitable transport mechanism.

[0034] Preferably, Client API 156 is an application programming interface used by management application 32 to communicate with object manager 20 using Remote Method Invocation (RMI) protocol 158 or XML over an HTTP protocol 160 according to the techniques described in U.S. patent application No. ________________ (Atty Docket SUN1P366) referenced above. Other suitable protocols may also be used such as COM from Microsoft Corporation. Client API 156 may communicate directly with object manager 20 using RMI or may communicate using the XML/HTTP protocol using web server 140. Alternatively, client API 156 can communicate using the XML/HTTP protocol 160 directly should object manager 20 support the HTTP format.

[0035] Connections 170a-170d are any suitable local or network connection between computer system 110 and application computer 150. By way of example, these connections occur over an internet, an intranet, an extranet, within a workgroup, or other.

[0036] FIG. 2B illustrates further detail in which CIM object manager 20 communicates with any of a variety of remote CIM repositories using a repository API 180. In this embodiment, each of repositories 190, 192 and 194 are located remotely from object manager 20 and computer system 110. In this example each repository is located on a different computer although it is conceivable that all may be located on a single computer, or that a repository is local to computer system 110.

[0037] Repository 190 is a database implemented using a flat file technique or object serialization in JAVA; it communicates with Repository API 180 over a network connection 132a that uses simple JAVA code protocol. Repository 192 is an object-oriented database and may be implemented using tools such as those available from Sybase, Oracle, or Informix. Repository 812 communicates with Repository API 180 over a network connection 132b using a JAVA Database Connectivity (JDBC) protocol. Repository 194 is a Lightweight Directory Access Protocol (LDAP) type of database that communicates over network connection 132c using a JAVA Naming Directory Interface (JNDI) protocol.

[0038] Repository API 180 is used by object manager 20 to store data to, or retrieve data from, the repositories. Repository API 180 includes a variety of classes, instances and methods useful for communicating with the repositories using any suitable protocol. Preferably, Repository API 180 communicates with the repositories using a JAVA language protocol, a JDBC protocol, a JNDI protocol, an LDAP protocol, an ODBC protocol, or other protocols suitable for use with a database. Implementation of such communication between object manager 20 and the repositories according to an embodiment of the invention is further described in FIGS. 4-8.

[0039] FIGS. 3A and 3B illustrate an example graphical user interface for CIM workshop 152. Preferably, a login to the workshop prompts for a host name,

namespace, user name and password. By default, workshop 152 connects to the object manager on the local host in the default namespace. FIG. 3A illustrates CIM classes that represent objects in the selected namespace on the selected host. Listed in panel 210 are the objects of the selected namespace. On the right-hand panel are shown the properties 212 for the selected object (in this case the object "Solaris Package") and methods 214 (not shown). FIG. 3B shows all instances of a selected object. Instances are shown in the left-hand panel, and in this example instance 252 is shown. The right-hand panel shows all properties 254 associated with the selected instance and its associated methods 256 (not shown).

[0040] FIG. 4 illustrates an interface definition 300 of Repository API 180 useful for implementing an embodiment of the invention. Interface 300 lists various methods that may be called by object manager 20 in the course of database operations with repositories 190-194. Advantageously, this interface may be implemented using a variety of classes having protocol-specific methods thus allowing a transport neutral object manager to be written. Interface 300 includes an interface name 302 which in this example is "CIM Repository API." Included are a variety of methods defined for the interface. Each method has a method name 304, a return value 306 and parameters 308. By way of example, shown is one method "Add CIM Element" 310 having a return value of "void" and accepting parameters "element" and "namespace." A large number of other methods may be defined for interface 300

[0041] By way of example, these methods include the following. The Create Namespace method creates a CIM namespace, a directory containing classes and instances. (When a management application connects to object manager 20 it specifies a namespace. All subsequent operations occur within that namespace on the object manager host.) The method Delete Class deletes the specified class. The method Delete Instance deletes the specified instance. The method Delete Qualifier deletes the specified qualifier. The method Enumerate Classes retrieves the specified classes from a repository. The method Enumerate Namespace gets a list of namespaces. The method Enumerate Instances gets of list of instances for the specified class. The method Enumerate Qualifier Types get a list of qualifier types for the specified class. The method Get Class gets the CIM class for the specified CIM object path. The Get Instance method gets the CIM instance for the specified CIM object path.

[0042] The method Get Qualifier Type gets the qualifier type for the specified CIM object path. The method Set Instance invokes a repository to add or update the specified CIM instance to the specified namespace. Other methods may also be included within interface 300 such as Add Aliased Class Name, Add Aliased Instance Name, Get Aliased Class Name, Get Aliased Instance Name, etc.

[0043] Once interface 300 has been defined it is possible to then code protocol-specific methods to implement each of the methods defined in interface 300. In this fashion, any number of protocol-specific classes are provided each having an implementation for a specific protocol such as JDBC or LDAP. Though the use of these protocol-specific classes, object manager 20 is able to communicate with any CIM repository using any suitable protocol in a transparent fashion.

[0044] FIG. 5 illustrates an implementation definition 400 for the interface of FIG. 4. Specifically, implementation 400 implements class "CIM Repository API" using methods specific to the LDAP protocol. Such protocol-specific methods allow object manager 20 to communicate via Repository API 180 to repository 194 using the LDAP protocol. Implementation 400 includes a class name 402 "CIM Repository LDAP." Also included is constructor definition code 404 that constructs an instance of class 402 that is specific to the LDAP protocol. Use of a constructor definition to create an instance is well known to those of skill in the art.

[0045] Also included in implementation 400 are the specific implementations of the methods defined upon interface 300. For each method implemented there is a method name 406, a return value 408, parameters 410 and implementation code 412. Implementation code 412 is preferably JAVA code that implements the particular method using any constructs necessary that are specific to the RMI protocol. Those of skill in the art will appreciate how to implement JAVA code for a particular purpose that must adhere to a specific protocol.

[0046] Preferable, all of the methods defined upon interface 300 are implemented in implementation 400. Shown by way of example is the method Add CIM Element 416 which has a return value of "void" and accepts the parameters element and namespace. Not shown for simplicity is the actual LDAP-specific JAVA code that implements the method Add CIM Element. The other methods defined in interface 300 are also listed in implementation 400 along with their LDAP-specific code.

[0047] FIG. 6 illustrates an implementation definition 500 for the interface of FIG. 4. Specifically, implementation 500 implements class "CIM Repository API" using methods specific to the JDBC protocol. Such protocol-specific methods allow object manager 20 to communicate via Repository API 180 to repository 192 using the JDBC protocol. Implementation 500 includes a class name 502 "CIM Repository JDBC." Also included is constructor definition code 504 that constructs an instance of class 502 that is specific to the JDBC protocol. Use of a constructor definition to create an instance is well known to those of skill in the art.

[0048] Also included in implementation 500 are the specific implementations of the methods defined upon interface 300. For each method implemented there is a method name 506, a return value 508, parameters 510 and implementation code 512. Implementation code

512 is preferably JAVA code that implements the particular method using any constructs necessary that are specific to the JDBC protocol. Those of skill in the art will appreciate how to implement JAVA code for a particular purpose that must adhere to a specific protocol.

[0049] Preferable, all of the methods defined upon interface 300 are implemented in implementation 500. Shown by way of example is the method Add CIM Element 516 which has a return value of "void" and accepts the parameters element and namespace. Not shown for simplicity is the actual JDBC-specific JAVA code that implements the method Add CIM Element. The other methods defined in interface 300 are also listed in implementation 500 along with their JDBC-specific code.

[0050] FIG. 7 is a JAVA factory class 600. Factory 600 is used for determining which protocol is desired by object manager 20 and directing the creation of a protocol-specific object to be returned to the object manager.

[0051] Factory 600 includes a class name 602 "CIM Repository Factory" and any number of defined methods. For each method there is a method name 604, a return value 606, parameters 608 and an implementation 610. In particular, the method Get Repository API accepts the parameters protocol, namespace and version, and returns an instance of interface 300 which is a protocol-specific instance of either implementation 400 or implementation 500. Of course, other protocol-specific objects may be returned if other implementations are defined. The implementation code 610 for method 612 may be any suitable JAVA code that checks the protocol parameter to see which protocol is desired and then directs either implementation 400 or 500 to construct a new instance of itself. By way of example, a series of case statements may be used. Other methods may also be defined and implemented within factory 600.

OBJECT MANAGER EXECUTION

[0052] FIG. 8 is a flowchart illustrating invocation of a method by the object manager to perform a database operation such as storing or retrieving an object. Once management application 32 has been created by developer 30 and the classes and methods of FIGS. 4-7 have been defined, the object manager may perform operations on one of the repositories using any desired protocol. FIG. 8 illustrates a single method call according to one embodiment of the invention.

[0053] In step 702 management application 32 creates a connection from application computer 150 to computer system 110. Preferably, application 32 invokes a method within Client API 156 which creates an instance of application 32 within object manager 20. Application 32 passes to the method a host name, a namespace, a user name, a password, and the protocol by which it is desired to communicate with host computer system 110. Any suitable network protocol may be identified such as RMI, XML/HTTP or DCOM.

[0054] In step 704 object manager 20 receives a method call from application 32 that requires a database operation. The method call is preferably performed using the technique described in U.S. patent application No. 09/333,878. In response to this method call, object manager 20 identifies a repository and protocol and makes a call to Repository API 180.

[0055] In step 706 factory 600 of Repository API 180 checks the protocol desired by object manager 20 using its method Get Repository API. This method returns a protocol-specific object which is an instance of either the class defined in implementation 400 or the class defined in implementation 500. For example, in step 710 if the protocol parameter is LDAP, then in step 714 the constructor definition 404 of implementation 400 executes and results in an LDAP-specific object having LDAP-specific methods being returned to the object manager. On the other hand, if the protocol parameter is JDBC, then in step 722 the constructor definition 504 of implementation 500 executes and produces a JDBC-specific object which is returned to the object manager. As shown in steps 726 and 730, a desire for use of simple JAVA protocol results in a JAVA-specific object being returned. Other protocols are also supported. In step 738 object manager 20 invokes a desired database method upon the protocol-specific object recently returned. Because the methods of this object are specific to the protocol desired by object manager 20, communication between Repository API 180 and the target repository occurs using the desired protocol in a fashion transparent to application 32 and to object manager 20.

[0056] Once the target CIM repository has processed the method (which may be a request for an object, a request to add an object, etc.), then in step 742 the result is returned from the repository to object manager 20 via Repository API 180 using the desired protocol. In this fashion, a technique has been described that allows an object manager to be written independent of the protocol by which it is desired to communicate with a target CIM repository.

COMPUTER SYSTEM EMBODIMENT

[0057] FIGS. 9 and 10 illustrate a computer system 900 suitable for implementing embodiments of the present invention. FIG. 9 shows one possible physical form of the computer system. Of course, the computer system may have many physical forms ranging from an integrated circuit, a printed circuit board and a small handheld device up to a huge super computer. Computer system 900 includes a monitor 902, a display 904, a housing 906, a disk drive 908, a keyboard 910 and a mouse 912. Disk 914 is a computer-readable medium used to transfer data to and from computer system 900.

[0058] FIG. 10 is an example of a block diagram for computer system 900. Attached to system bus 920 are

a wide variety of subsystems. Processor(s) 922 (also referred to as central processing units, or CPUs) are coupled to storage devices including memory 924. Memory 924 includes random access memory (RAM) and read-only memory (ROM). As is well known in the art, ROM acts to transfer data and instructions uni-directionally to the CPU and RAM is used typically to transfer data and instructions in a bi-directional manner. Both of these types of memories may include any suitable of the computer-readable media described below. A fixed disk 926 is also coupled bi-directionally to CPU 922; it provides additional data storage capacity and may also include any of the computer-readable media described below. Fixed disk 926 may be used to store programs, data and the like and is typically a secondary storage medium (such as a hard disk) that is slower than primary storage. It will be appreciated that the information retained within fixed disk 926, may, in appropriate cases, be incorporated in standard fashion as virtual memory in memory 924. Removable disk 914 may take the form of any of the computer-readable media described below.

[0059] CPU 922 is also coupled to a variety of input/output devices such as display 904, keyboard 910, mouse 912 and speakers 930. In general, an input/output device may be any of: video displays, track balls, mice, keyboards, microphones, touch-sensitive displays, transducer card readers, magnetic or paper tape readers, tablets, styluses, voice or handwriting recognizers, biometrics readers, or other computers. CPU 922 optionally may be coupled to another computer or telecommunications network using network interface 940. With such a network interface, it is contemplated that the CPU might receive information from the network, or might output information to the network in the course of performing the above-described method steps. Furthermore, method embodiments of the present invention may execute solely upon CPU 922 or may execute over a network such as the Internet in conjunction with a remote CPU that shares a portion of the processing.

[0060] In addition, embodiments of the present invention further relate to computer storage products with a computer-readable medium that have computer code thereon for performing various computer-implemented operations. The media and computer code may be those specially designed and constructed for the purposes of the present invention, or they may be of the kind well known and available to those having skill in the computer software arts. Examples of computer-readable media include, but are not limited to: magnetic media such as hard disks, floppy disks, and magnetic tape; optical media such as CD-ROMs and holographic devices; magneto-optical media such as floptical disks; and hardware devices that are specially configured to store and execute program code, such as application-specific integrated circuits (ASICs), programmable logic devices (PLDs) and ROM and RAM devices. Examples of computer code include machine code, such as produced by a compiler, and files containing higher level code that are executed by a computer using an interpreter.

[0061] Although the foregoing invention has been described in some detail for purposes of clarity of understanding, it will be apparent that certain changes and modifications may be practiced within the scope of the appended claims. For instance, the application computer and the computer system to be managed may be the same computer, or may be separated by a great distance. Also, the various CIM repositories may be located along with the computer system, may each be remotely located on a separate computer, or may be remotely located on a single computer. The use of a web server may not be required should the CIM object manager support the HTTP format. Other types of classes and methods may be used while not departing from the spirit of the invention. Therefore, the described embodiments should be taken as illustrative and not restrictive, and the invention should not be limited to the details given herein but should be defined by the following claims and their full scope of equivalents.

Claims

1. A method for communication between a Common Information Model (CIM) object manager of a host computer and a CIM repository, said method comprising:

    creating a connection between said object manager and said CIM repository;

    passing a protocol indicator from said object manager to a repository application programming (API), said protocol indicator identifying a protocol by which said CIM object manager desires to communicate with said CIM repository;

    creating a protocol-specific object having methods implemented using said protocol; and

    returning said protocol-specific object to said CIM object manager, whereby said CIM object manager may communicate with said CIM repository using said protocol.

2. The method of claim 1 further comprising:

    invoking a method defined upon said protocol-specific object;

    transmitting said method using said protocol over said connection to said CIM repository; and

returning a result to said CIM object manager over said connection using said protocol.

3. The method of claim 1 wherein said protocol is LDAP, JDBC, or JAVA.

4. The method of claim 1 wherein said CIM repository is resident on said host computer.

5. The method of claim 1 wherein said CIM repository is resident on a separate computer.

6. The method of claim 1 wherein said creating a protocol-specific object includes

calling a JAVA factory class.

7. A computer system for interacting with a CIM repository database, said system comprising:

a CIM object manager including a protocol indicator and program code for interacting with said CIM repository; and

a repository application programming interface (repository API) including

a factory class arranged to receive said protocol indicator from said object manager and produce a protocol-specific object,

a first class having methods defined thereon implemented in a first protocol, and

a second class having methods defined thereon implemented in a second protocol, whereby said protocol-specific object may be returned to said object manager for use in interacting with said CIM repository.

8. The system of claim 7 wherein said CIM object manager is arranged to receive a method call from a management application using the protocol identified by said protocol indicator.

9. The system of claim 7 wherein said CIM repository is resident on said computer system.

10. The system of claim 7 wherein said computer system and said CIM repository are connected over a network connection implemented in the protocol identified by said protocol indicator.

11. The system of claim 7 wherein the protocol identified by said protocol indicator is LDAP, JDBC or JAVA.

12. The system of claim 7 further comprising:

a plurality of CIM repositories, each repository arranged to communicate with said CIM object manager using a different protocol.

13. The system of claim 12 wherein each repository is resident on a different computer.

14. A computer-readable medium comprising computer code for communication between a Common Information Model (CIM) object manager of a host computer and a CIM repository, said computer code of said computer-readable medium effecting the following:

creating a connection between said object manager and said CIM repository;

passing a protocol indicator from said object manager to a repository application programming (API), said protocol indicator identifying a protocol by which said CIM object manager desires to communicate with said CIM repository;

creating a protocol-specific object having methods implemented using said protocol; and

returning said protocol-specific object to said CIM object manager, whereby said CIM object manager may communicate with said CIM repository using said protocol.

15. The computer-readable medium of claim 14 further comprising computer code for effecting the following:

invoking a method defined upon said protocol-specific object;

transmitting said method using said protocol over said connection to said CIM repository; and

returning a result to said CIM object manager over said connection using said protocol.

16. The computer-readable medium of claim 14 wherein said protocol is LDAP, JDBC or JAVA.

17. The computer-readable medium of claim 14 wherein said creating a protocol-specific object includes

calling a JAVA factory class.

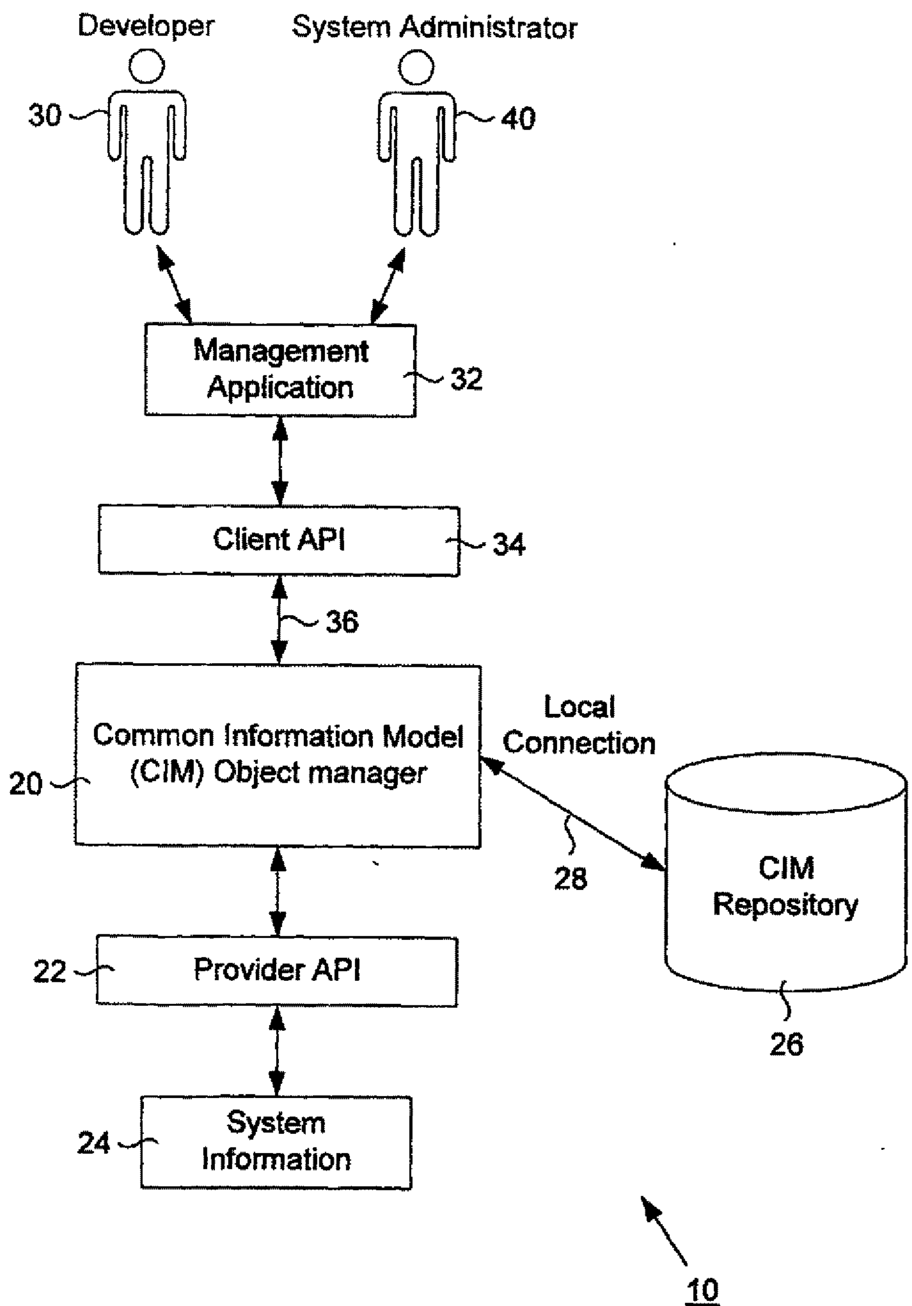


FIG. 1
(Prior Art)

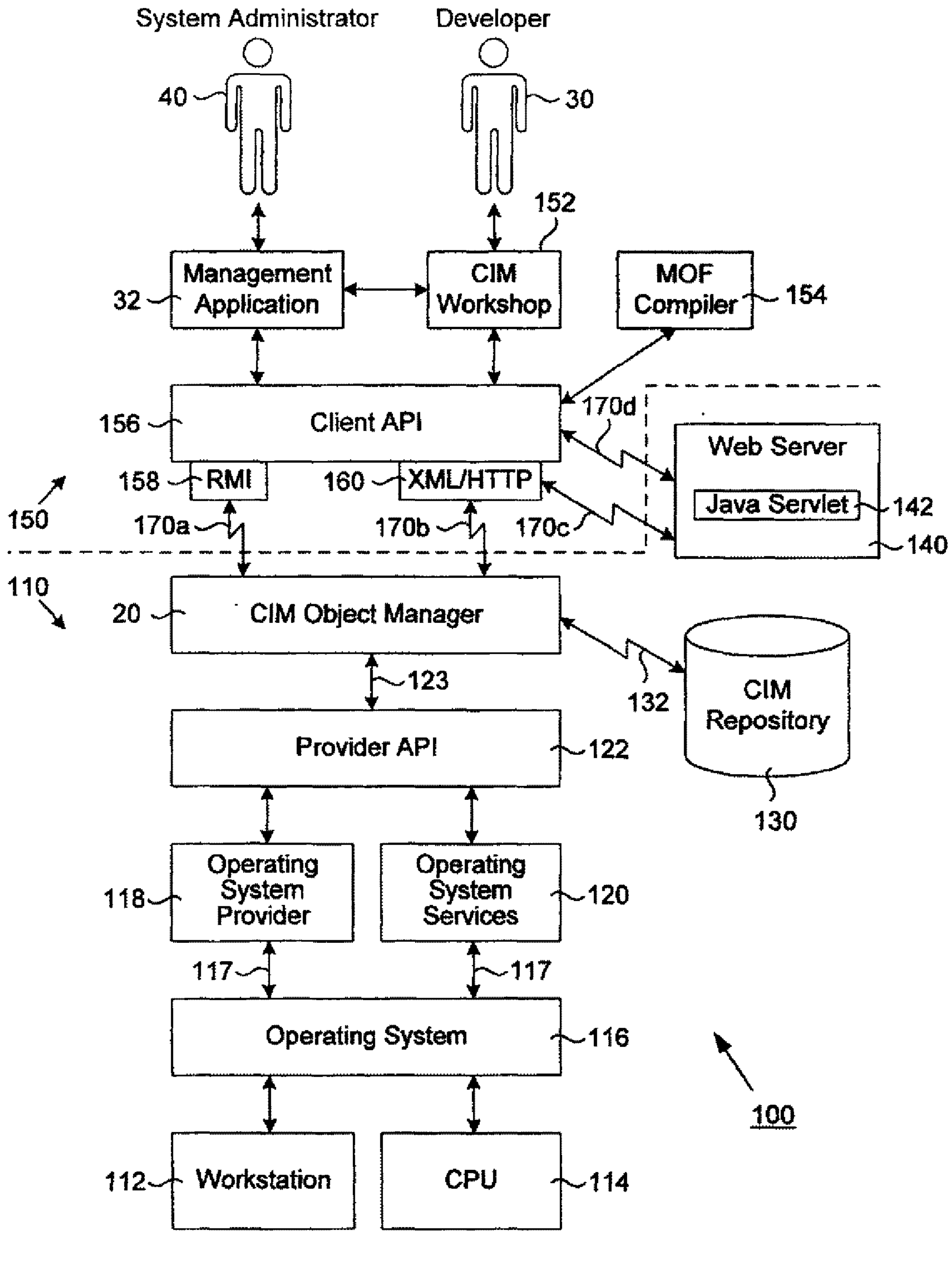

System Administrator
Developer
40
30
152
Management Application
32
CIM Workshop
MOF Compiler
154
Client API
156
170d
Web Server
RMI
158
160
XML/HTTP
Java Servlet
142
150
170a
170b
170c
140
110
20
CIM Object Manager
123
132
CIM Repository
Provider API
122
130
Operating System Provider
118
Operating System Services
120
117
117
Operating System
116
100
Workstation
112
CPU
114

FIG. 2A

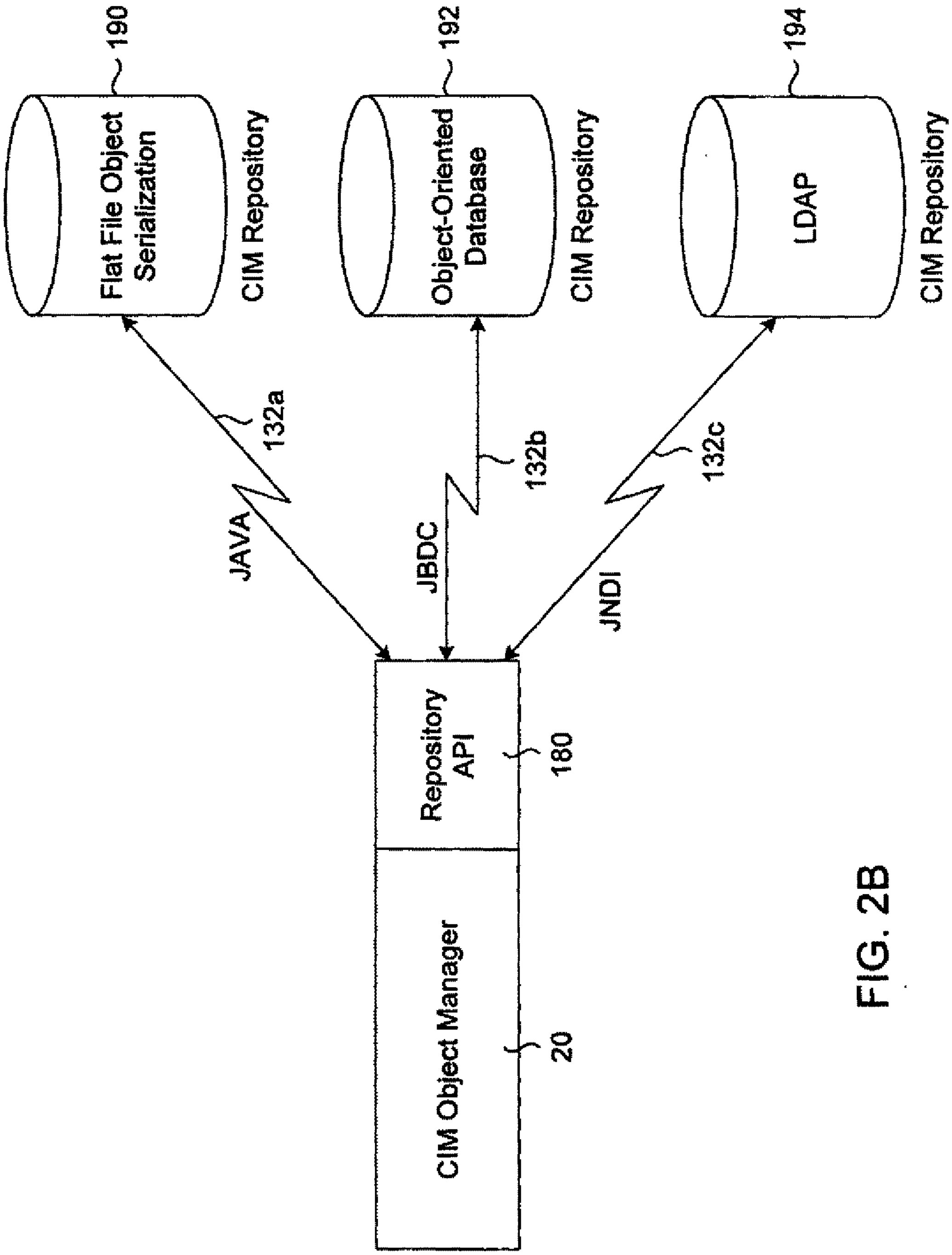

Flat File Object Serialization
190
CIM Repository
Object-Oriented Database
192
CIM Repository
LDAP
194
CIM Repository
132a
JAVA
132b
JBDC
132c
JNDI
Repository API
180
CIM Object Manager
20

FIG. 2B

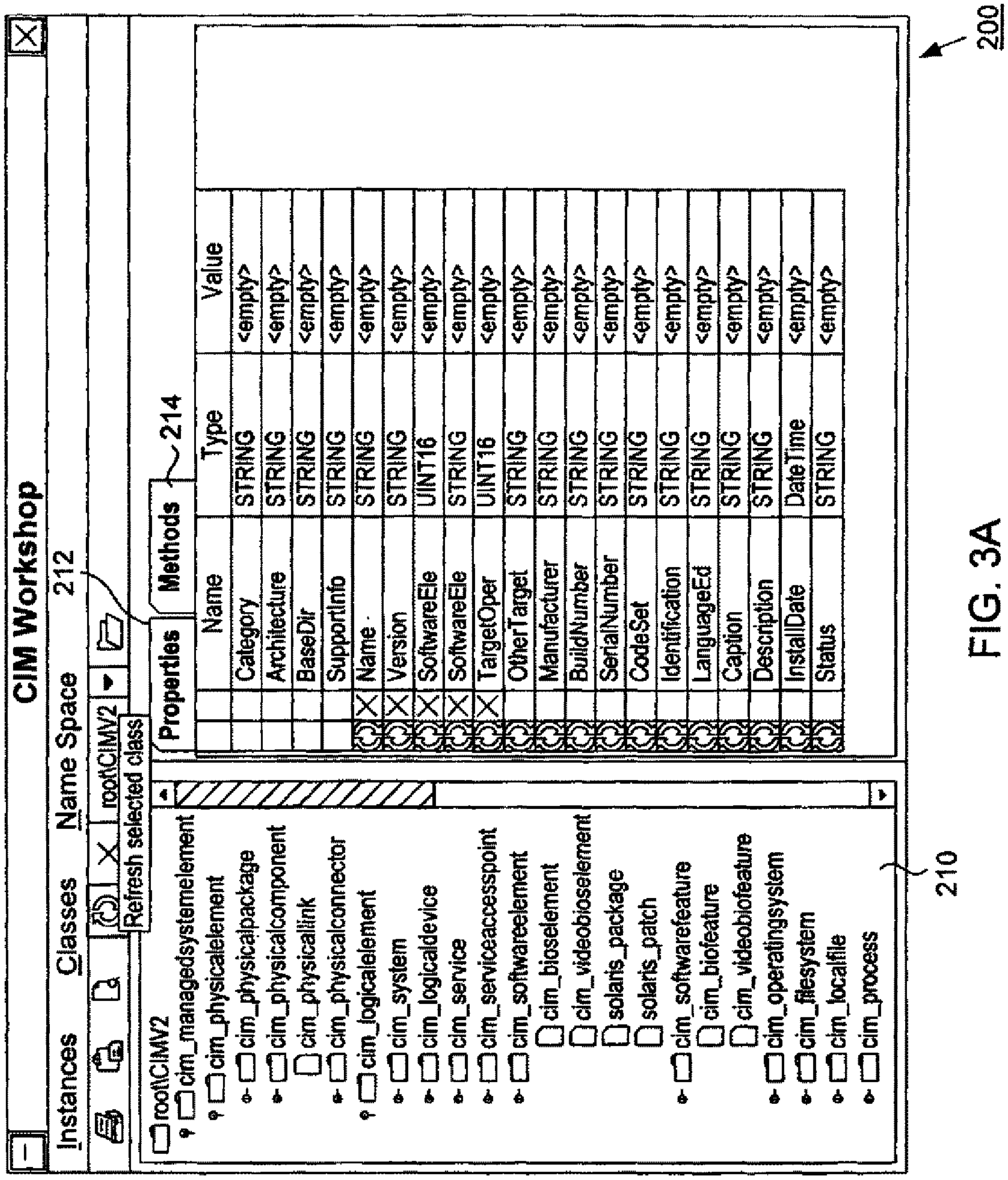

CIM Workshop
Instances
Classes
Name Space
212
root\CIMV2
Refresh selected class
root\CIMV2
cim_managedsystemelement
cim_physicalelement
cim_physicalpackage
cim_physicalcomponent
cim_physicallink
cim_physicalconnector
cim_logicalelement
cim_system
cim_logicaldevice
cim_service
cim_serviceaccesspoint
cim_softwareelement
cim_bioselement
cim_videobioselement
solaris_package
solaris_patch
cim_softwarefeature
cim_biofeature
cim_videobiofeature
cim_operatingsystem
cim_filesystem
cim_localfile
cim_process
Properties
Methods
214
Name
Type
Value
Category STRING <empty>
Architecture STRING <empty>
BaseDir STRING <empty>
SupportInfo STRING <empty>
Name STRING <empty>
Version STRING <empty>
SoftwareEle UINT16 <empty>
SoftwareEle STRING <empty>
TargetOper UINT16 <empty>
OtherTarget STRING <empty>
Manufacturer STRING <empty>
BuildNumber STRING <empty>
SerialNumber STRING <empty>
CodeSet STRING <empty>
Identification STRING <empty>
LanguageEd STRING <empty>
Caption STRING <empty>
Description STRING <empty>
InstallDate DateTime <empty>
Status STRING <empty>
210
200

FIG. 3A

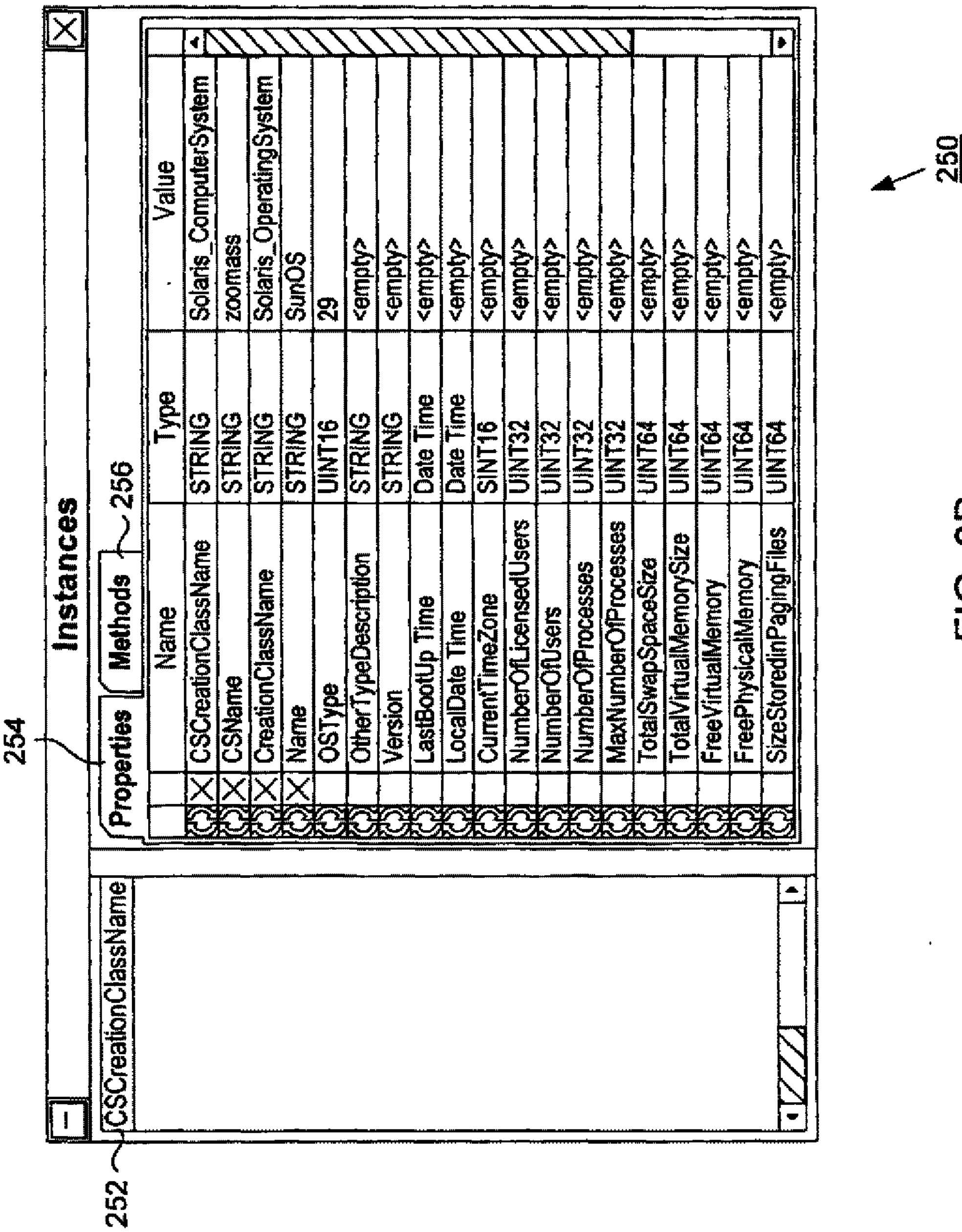
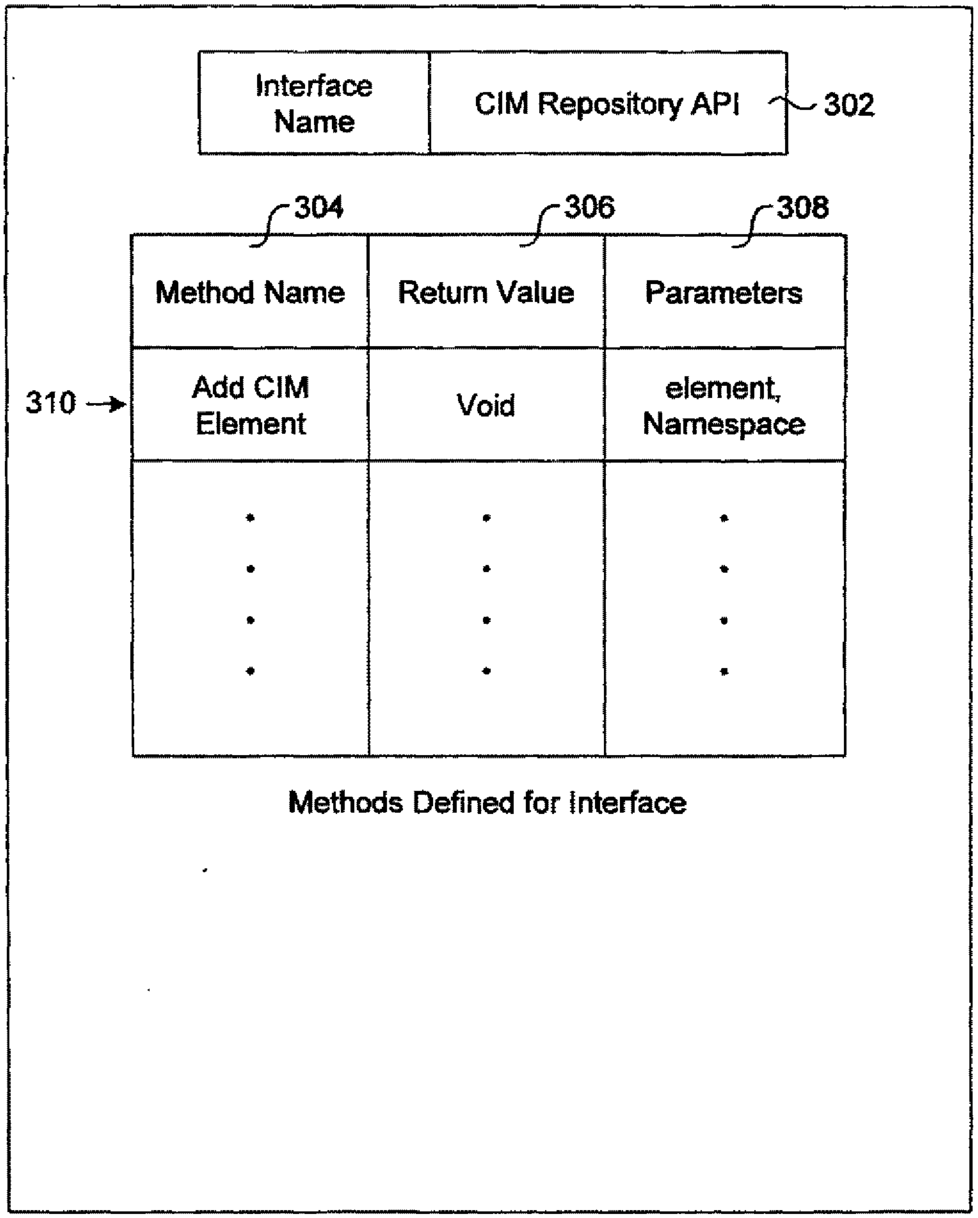

Instances
CSCreationClassName
252
254
256
Properties
Methods
Name
Type
Value
CSCreationClassName
STRING
Solaris_ComputerSystem
CSName
STRING
zoomass
CreationClassName
STRING
Solaris_OperatingSystem
Name
STRING
SunOS
OSType
UINT16
29
OtherTypeDescription
STRING
<empty>
Version
STRING
<empty>
LastBootUp Time
Date Time
<empty>
LocalDate Time
Date Time
<empty>
CurrentTimeZone
SINT16
<empty>
NumberOfLicensedUsers
UINT32
<empty>
NumberOfUsers
UINT32
<empty>
NumberOfProcesses
UINT32
<empty>
MaxNumberOfProcesses
UINT32
<empty>
TotalSwapSpaceSize
UINT64
<empty>
TotalVirtualMemorySize
UINT64
<empty>
FreeVirtualMemory
UINT64
<empty>
FreePhysicalMemory
UINT64
<empty>
SizeStoredinPagingFiles
UINT64
<empty>
250

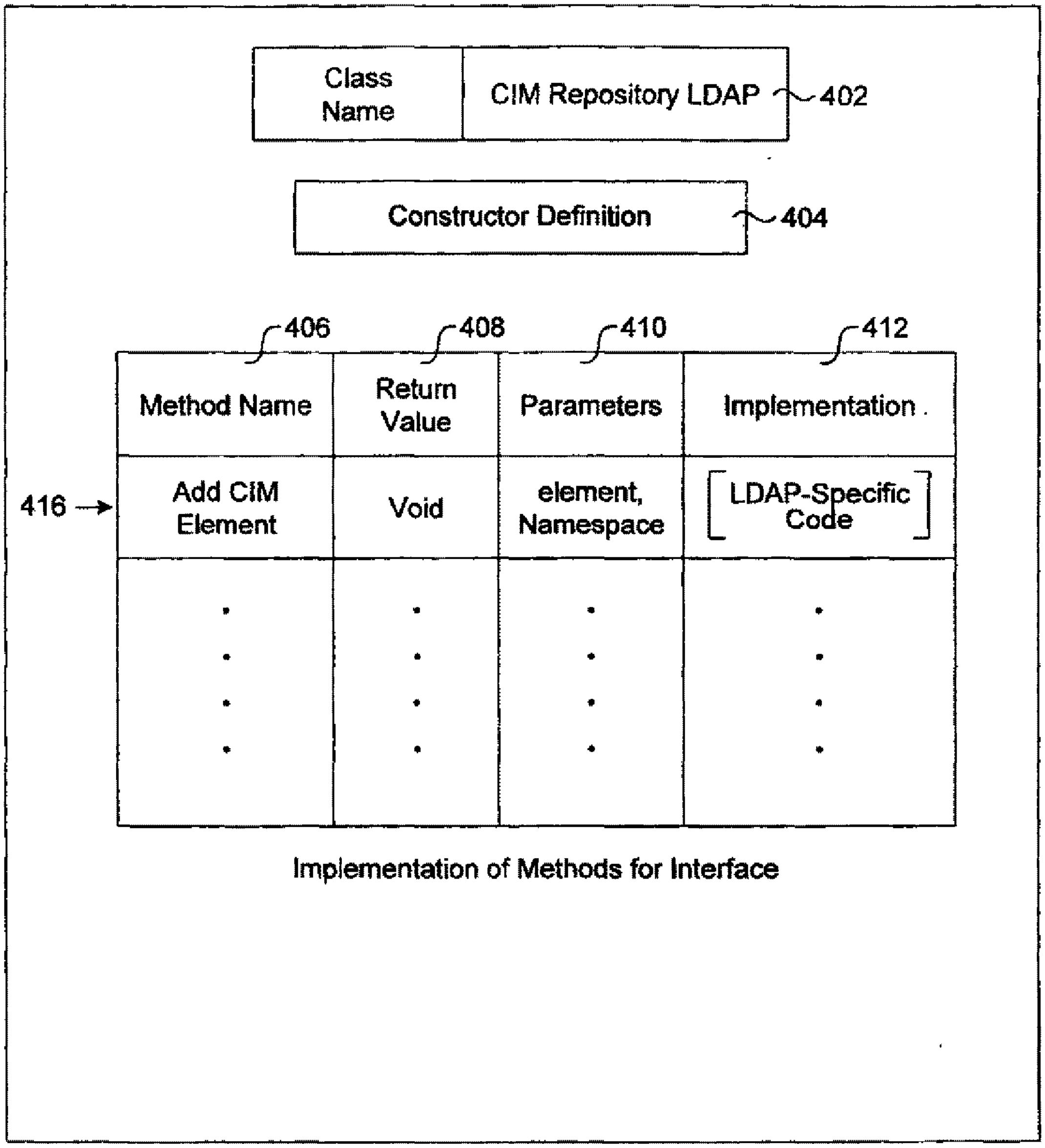
FIG. 3B



Interface Definition

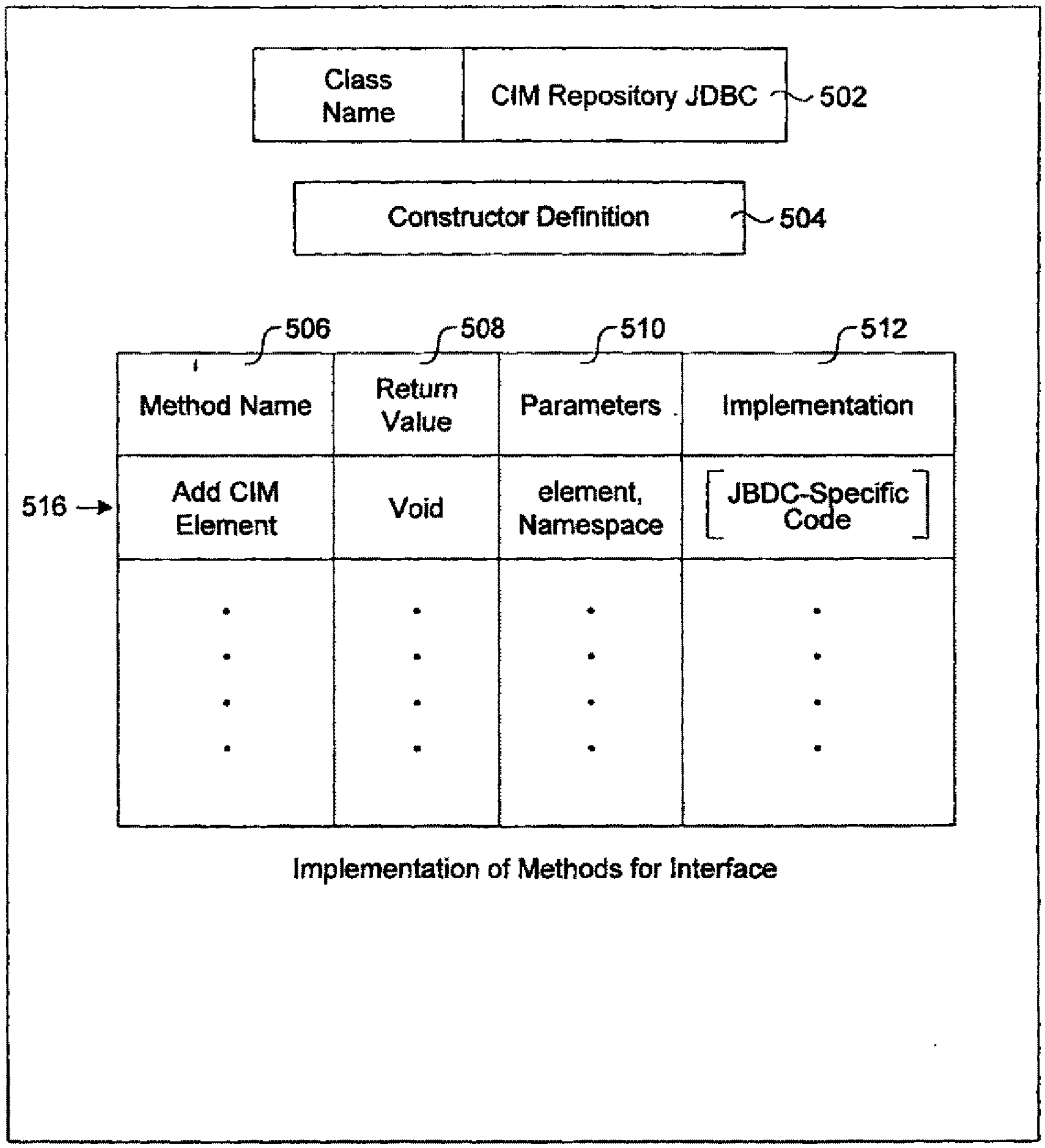
300

FIG. 4



Implementation Definition for Interface

FIG. 5



Implementation Definition for Interface

500

FIG. 6

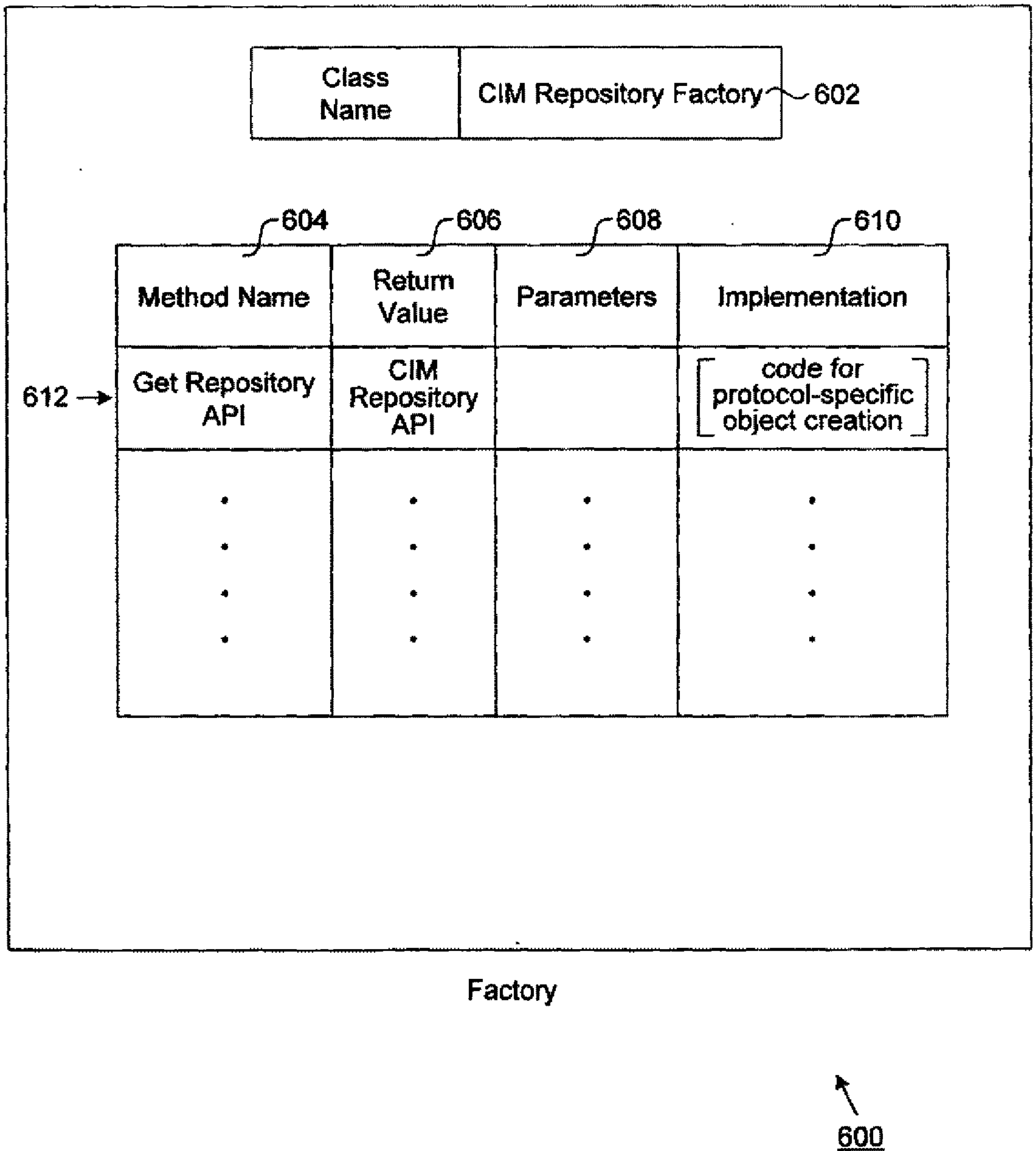

Class Name
CIM Repository Factory
602
604
606
608
610
Method Name
Return Value
Parameters
Implementation
612
Get Repository API
CIM Repository API
code for protocol-specific object creation
Factory
600

FIG. 7

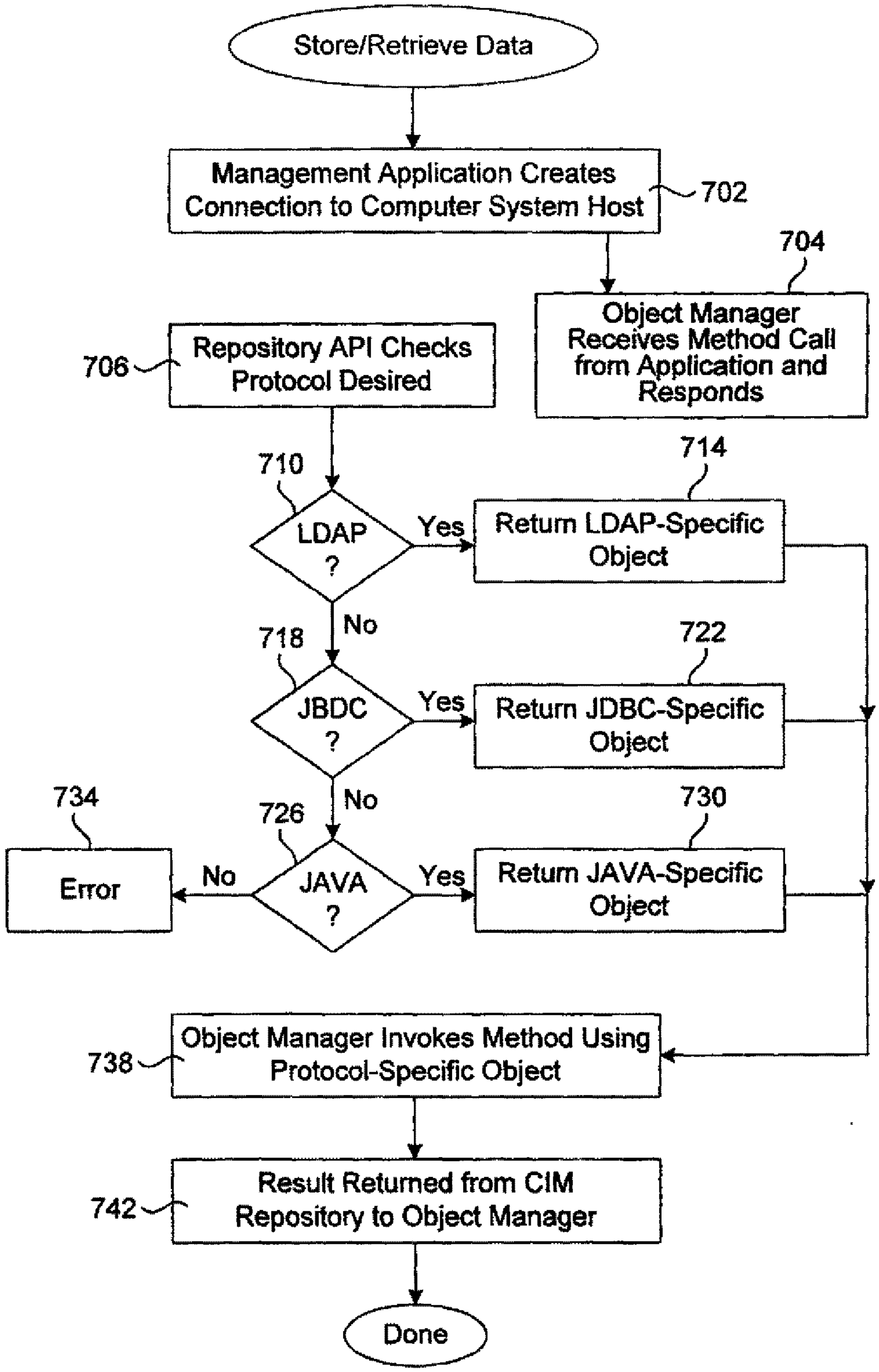

Store/Retrieve Data
Management Application Creates Connection to Computer System Host
702
704
Object Manager Receives Method Call from Application and Responds
706
Repository API Checks Protocol Desired
710
LDAP ?
Yes
Return LDAP-Specific Object
714
No
718
JBDC ?
Yes
Return JDBC-Specific Object
722
No
726
JAVA ?
Yes
Return JAVA-Specific Object
730
No
Error
734
738
Object Manager Invokes Method Using Protocol-Specific Object
742
Result Returned from CIM Repository to Object Manager
Done

FIG. 8

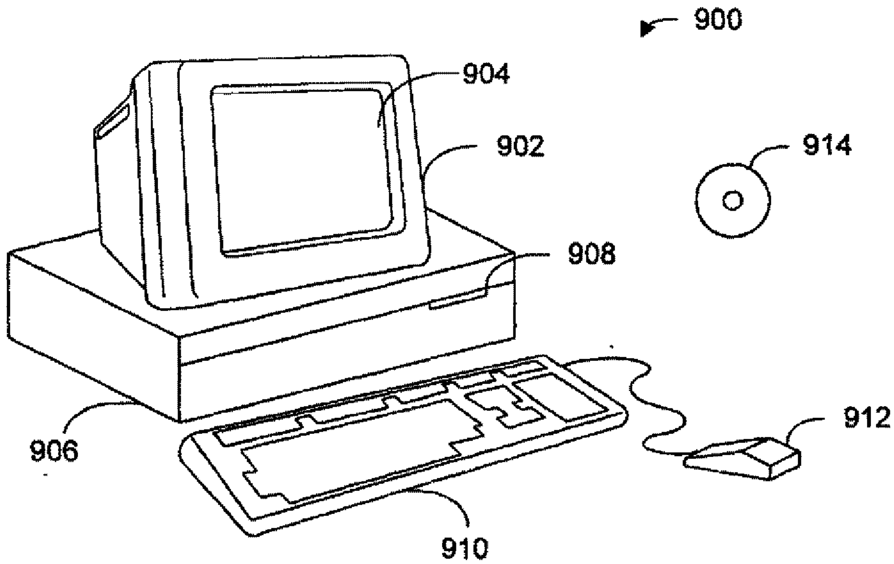

900
904
902
914
908
906
912
910

FIG. 9

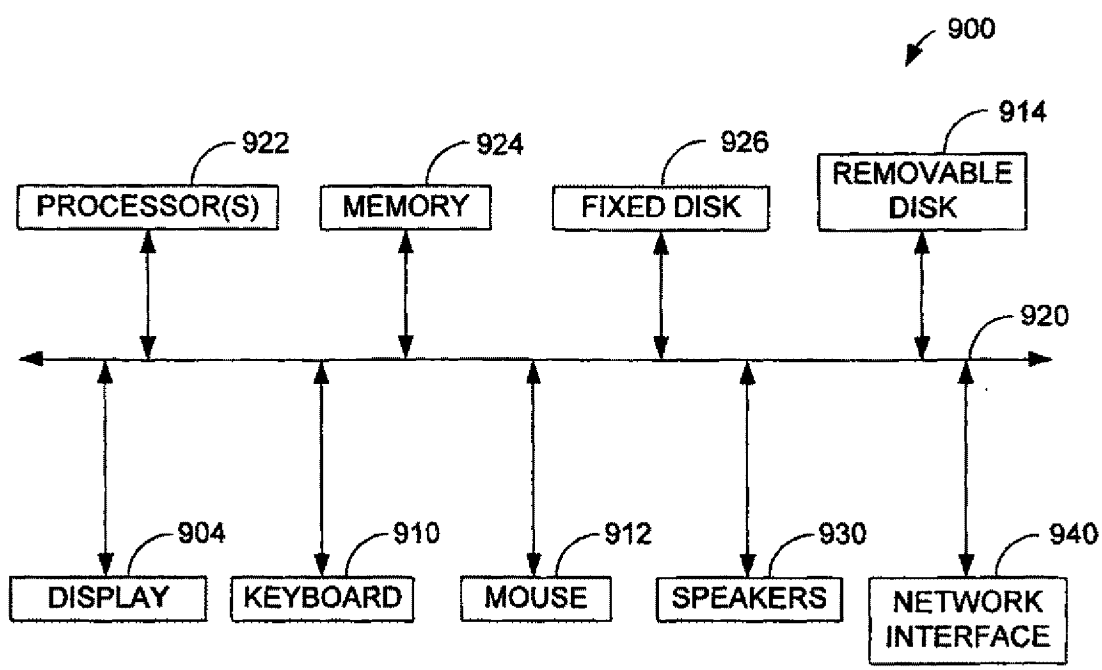

900
922
924
926
914
PROCESSOR(S)
MEMORY
FIXED DISK
REMOVABLE DISK
920
904
910
912
930
940
DISPLAY
KEYBOARD
MOUSE
SPEAKERS
NETWORK INTERFACE

FIG. 10

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(54) Method of managing time-slot interchange in transoceanic MS-spring networks

(57) A method of managing time slot interchange in transoceanic MS-SP ring networks. The method, in case of ring failure in a single span of the path installed in a transoceanic MS-SP RING with Time Slot Interchange (TSI) mechanism, comprises the step of carrying out a ring switch action by the MS-SP mechanism, and is characterized by comprising the step of re-routing the path in the time slot of the low-priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the failured span. The method according to the invention further provides for the managing of double-failure or multi-failure cases resulting in one or more nodes being isolated.

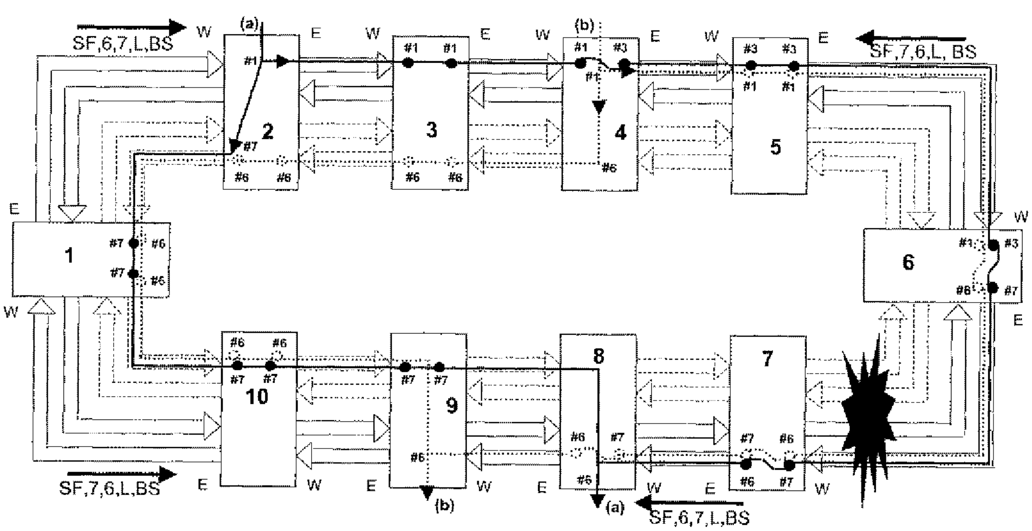


Fig. 3

EP 1 185 015 A2

Description

[0001] The present invention relates to a method of managing changes of time-slot allocations in ring networks protected by a transoceanic MS-SPRING protection mechanism.

[0002] In SDH MS-SPRING (Multiplex Section Shared Protection Ring) networks, a shared protection mechanism is implemented, which mechanism allows for the automatic restoration of the traffic in case of defects or failures in the connection fibers. In other words, the MS shared protection ring networks perform the automatic restoration of the traffic through a synchronized re-routing of said traffic, performable at each node of the ring. This operation is controlled by a protocol consisting of messages that are continuously interchanged between adjacent nodes. Said protocol and the related operations are defined by several international standards issued by ANSI, ITU-T and ETSI and they are characterized by a set of rules and messages. See, for instance, Recommendation ITU-T G. 841.

[0003] Protection in an MS shared protection ring network is implemented according to a so-called Bridge and Switch technique that consists essentially in re-routing the traffic, by means of an appropriate modification in the internal connections of the network elements, switching it from the working capacity to the protection capacity. The protection in an MS shared protection ring network is a multiplex section-oriented protection mechanism, i.e. the events defining the traffic restoration and the hierarchy that regulates those events are given at multiplex section level. In the "classic" (or terrestrial) MS shared protection rings, in the event of a failure, the whole high-priority line capacity is re-routed by utilizing the corresponding low-priority line capacity; in the transoceanic MS shared protection rings, on the contrary, only the paths affected by a failure are selectively re-routed.

[0004] It is also known that the ring networks provide for a mechanisms termed "Time Slot Interchange", in short TSI. TSI means, for instance, that when traffic is configured in a given ring network, such a traffic, which is carried in the associated STM-n and hence in the AU-4 contained in the STM-n, is allowed to travel through a network element occupying different AU-4 numbers at the input and at the output. Consider for instance a maximum capacity of a four-fiber ring, composed of sixteen AU-4s. The TSI mechanism allows one to enter a network element (of pure transit and where no termination occurs) with AU-4#X from its West side (W) and to go out from its East side (E) with an AU-4#Y, where X≠Y = 1, 2, ..., 16. The advantage is a greater flexibility in the traffic allocation and a very efficient exploitation of the band.

[0005] At present, performing TSI in ring networks protected by an MS-SPRING protection mechanism is not known. In particular, it is not known to perform allocation changes in transoceanic MS shared protection ring networks.

[0006] Therefore the main object of the present invention is to provide a method allowing the execution of allocation changes in transoceanic rings protected by an MS-SPRING mechanism. This and further objects are achieved by a method having the features set forth in independent claim 1 and a network element according to claim 8. The respective dependent claims define further advantageous characteristics of the invention itself. All the claims are intended to be an integral part of the present description.

[0007] The basic idea of the present invention consists substantially in protecting the high-priority traffic by assigning, in case of a ring failure, the low-priority channel time slots chosen according to the real failure location and to the instant at which such failure has occurred, with respect to other failures possibly already present.

[0008] The invention will certainly become clear in view of the following detailed description, given by way of a mere non limiting and exemplifying example, wherein:

- Fig. 1 shows a ring network in a stable faultless situation, the network having a plurality of nodes, two installed paths and some allocation changes;
- Fig. 2 shows the same ring network of Fig. 1 just after a ring failure took place;
- Fig. 3 shows the ring network of Fig. 2 in a stable situation with a ring failure;
- Figs. 4 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of simultaneous double failure;
- Figs. 5 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of nearly simultaneous double failure;
- Figs. 6 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of double failure at different times (first sub-scenario);
- Figs. 7 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of double failure at different times (second sub-scenario);
- Figs. 8 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of clearing of a first failure; and
- Figs. 9 show the signaling that is received/generated by the single nodes and the corresponding actions performed in the event of clearing of a second failure.

[0009] In the various figures, a four-fiber transoceanic telecommunication ring has been always depicted. The two working fibers (otherwise known as "high-priority channels" or "HP channels") are indicated by solid-line

arrows whereas the protection fibers (otherwise known as "low-priority channels" or "LP channels") are indicated by dashed-line arrows. Naturally, the present invention applies both to the illustrated case of bi-directional traffic and to the case of unidirectional traffic.

[0010] Moreover, the present invention is applicable also to rings in which the traffic subjected to **TSI** is configured with "channel concatenation (AU4)".

[0011] The ring illustrated to describe the invention comprises ten network elements or nodes, represented by blocks and designated by respective numerals (1 to 10). The West (W) and East (E) sides of each node are indicated. The term "span" is used throughout this description to mean that part between two adjacent nodes, for instance between nodes 1 and 2 or the one between nodes 7 and 8.

[0012] In the ring there are depicted, by way of a non-limiting example, two paths installed, "path (a)" and "path (b)". The first path (path a) is depicted by a bolt solid line whereas the second path (path b) is depicted by a bolt dotted line. Path (a) is inserted at node 2 and is dropped at node 8. Path (b) is inserted at node 4 and is dropped at node 9.

[0013] Finally, it has been tried to clearly indicate (with numbers after symbol "#") the various time slots in which the various paths are allocated, span by span. Thus it has been also indicated if a Time Slot Interchange (TSI) occurs at a node or if that node allows that path to transit without changing the AU-4 on which it is allocated.

[0014] The present invention contemplates the general criteria set forth below:

I) single failure: once a ring failure has occurred in a given span, a ring switch action is performed by the MS shared protection mechanism. This activity defines the set of re-routable paths, namely all the paths whose nominal route includes the failured span. According to the present invention, each of these paths is re-routed on the time slot of the low-priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the failured span. There is no risk of any conflict since the LP time-slot assignment criterion is the same for all the failured paths.

II) double failure: if a failure occurs at a further span and the path can still be saved, then

II.I) i) the actual re-routing is released; ii) among the two failured spans one is chosen according to a certain criterion; and iii) the path is re-routed over the time slot of the low-priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the selected failured span. Should multiple (more than two) failures occur, the choice of the span to consider for the **TSI** path protection is to be made by selecting, according to the above criterion, among the spans adjacent to the switching nodes that are able to communicate with the termination nodes of the path to be protected. There is no risk of any conflict since the LP time-slot assignment criterion is the same for all the paths affected by the failure. In this way, any transient misconnection is avoided.

II.II) The actual re-routing is not released when the persistency of the re-routing information is supported by the ring network elements.

[0015] The procedures that are implemented by each node of the ring (in addition to the procedures already provided for by the MS shared protection mechanism) will be indicated below:

A. If, at both W and E sides of the node, two Bridge Requests with an "Idle" status code concerning the same span (single failure) are detected, then each path comprising the span in question is re-routed over the LP time slot equal to the HP time slot of the (sole) span affected by the failure. In other words, if the node is a path termination node (a node where the path to be re-routed is inserted or dropped), the Bridge & Switch action is performed by utilizing the LP time slot equal to the HP time slot of the span affected by the failure.

A1. If at W or E side of the node at least one Bridge Request with an "Idle" status code is detected, a pass-through is performed, if necessary, by utilizing the same LP time slot (LP time slot equal to the HP time slot of the span affected by the failure).

B. If at both W and E sides of the node Ridge Requests with a "Bridged and Switched" status code concerning different spans (multiple failures) are detected, then each path comprising the spans in question is re-routed over the LP time slot equal to the HP time slot of the span affected by the failure that has been chosen according to a fixed criterion. The criterion for selecting one among the failed spans could, for instance, be any one of the following:

a) the failed span adjacent to the switching node with higher (or lower) node identification is chosen;

b) the failed span adjacent to the switching node coming first (or last) in the ring map; or

c) the failed span adjacent to the switching node that is "far West" or "far East" node in the ring.

[0016] Similarly to the above case A, if the node in question is a path termination node (node in which the path to be re-routed is inserted or dropped), the Bridge & Switch action is performed by utilizing the LP time slot equal to the HP time slot of the selected failured span.

B1. If at the W or E side of the node at least one

Bridge Request with "Bridged and Switched" status code is detected, a pass-through is performed, if necessary, by utilizing the same LP time slot as above.

C. In an optimized embodiment, should a path re-routing due to single failure be under way, if at W and E sides of the node Bridge Requests with a different ("Idle" or "Bridged and Switched") status code that carry the indication of a second failure, hence located at a different span, are detected, then, for each path that has already been re-routed, it is evaluated if the new requests exhibit a failured state that requires to release or maintain the present re-routing. It is not necessary to release the actual re-routing of a single path in the following cases:

i) when a failure is detected in addition to the already existing failure/s (and such new failure/s does/do not result in the "isolation" condition of any of the termination nodes of the already protected path); or ii) when the last occurred span failure has been removed.

[0017] It is understood how the persistency of the path re-routing condition is strictly connected to the coexistence of the aforesaid triggers at both sides of the nodes. This behavior results in the correct re-routing of the failured paths, namely it prevents misconnections from being created during transient states of the ring protection mechanism.

[0018] Note that the above is applicable to the case of bi-directional path and unidirectional one not using the inverse route. Clearly, if two unidirectional paths are allocated on the same time slot (in opposite directions), then the same LP time slot can be assigned to both paths.

[0019] Fig. 1 shows a transoceanic MS shared protection ring protected against failures in accordance with the invention, at some nodes of the ring Time-Slot Interchange (TSI) occurring. The installed paths are two: (a) and (b). Path (a) enters the ring at node 2 and is allocated on the AU-4 #1; therefore, in span 2-3 the allocation is AU-4#1; at node 3 the allocation is unchanged (therefore it remains AU-4#1); at node 4, the allocation is changed from AU-4#1 to AU-4#3; in span 4-5 the allocation is therefore AU-4#3; at node 5 the allocation is not changed (therefore it remains AU-4#3); in span 5-6 the allocation is therefore AU-4#3; at node 6 the allocation is changed from AU-4#3 to AU-4#7; therefore, in span 6-7 the allocation is AU-4#7; at node 7 the allocation changes again from AU-4#7 to AU-4#6; therefore, in span 7-8 the allocation is AU-4#6; finally, path (a) is dropped at node 8. For path (b): it enters at node 4 and is allocated on AU-4#1; this allocation is maintained up to node 6 where it changes from AU-4#1 to AU-4#6; it is changed again at node 7 (passing from AU-4#6 to AU-4#7) and at node 8 (passing from AU-4#7 to AU-4#6). Finally, path (b) is dropped at node 9.

[0020] In the event of a ring failure (namely, a failure that makes both high-priority (HP) channels and low-priority (LP) channels useless), the present invention provides for a method of remedying such failure notwithstanding the presence of allocation changes in the ring. Reference should be made, for the event of single ring failure, to Figs 2 and 3 where a ring failure in the span 6-7 has been simulated.

[0021] As it is known, the management of failures in the synchronous (SDH or SONET) telecommunications networks occurs, for some protection types among which the MS-SPRING one, through bytes K1 and K2 of the frame overhead. Since the present invention does not concern specifically such bytes K1 and K2, a more precise description thereof will not be given, the reader having the possibility to refer to relevant Recommendations, for instance the ITU-T Recommendation G. 841, incorporated herein by reference.

[0022] In the event of a failure, the nodes (6 and 7) that are adjacent to the failure will send, as usual, proper failure signaling in the direction opposite to the failure. The structure of the request (APS signaling) is the following: Bridge Request, Destination Node ID, Source Node ID, Type of Path, Protection Status. In this instance, the node 6 will send a signaling of the type SF, 7,6,L,ID (Signal Fail, Destination Node: 7, Source Node: 6, Path: Long, Protection Status: Idle) to indicate that a ring failure occurred at span 6-7 and that no action has been taken for the time being. Node 7 will do the same by sending a signaling type SF,6,7,L,ID from its East side (E).

[0023] Such signaling will travel down the ring in opposite directions and will be received by termination nodes (2, 8; 4, 9) of paths (a) and (b) that will perform the requested Bridge and Switch (BR & SW) by utilizing the LP channels. In accordance with the present invention, the ring protection (BR&SW and pass-through), in the event of single ring failure, is performed by squelching the terminations of LP traffic allocated on the AU-4s corresponding to the failured span and by allocating the HP traffic on such AU-4s. With reference to Fig. 3, since in span 6-7 the path (a) was allocated on the HP AU-4#7 and the path (b) was allocated on the HP AU-4#6, the allocation AU-4#7 (of the LP channels) will be utilized for the first path and the allocation AU-4#6 (of the LP channels) will be utilized for the second path.

[0024] Should TSI be configured also on low-priority traffic, the high-priority traffic protection that requires the use of one of the LP channels utilized in the low-priority TSI, will anyway result in the squelching action on both the low-priority traffic terminations.

[0025] Once a node adjacent to the failure has received the signaling sent by its homologous opposite side, with Protection Status corresponding to "Idle", the node itself will send a modified signaling (with Protection Status = Bridged & Switched, BS). In other words, node 6 will send SF,7,6,L,BS from its West side whereas node

7 will send SF,6,7,L,BS from its East side. Upon restoration of the full ring functionality (fault clearing) the BR&SW will be removed and the failure signaling (SF, 7,6,L,BS and SF,6,7,L,BS) will be removed.

[0026] The present invention, in addition to the single failure event illustrated above, provides for a traffic protection method applicable to multiple failures leading to isolation of one or more nodes in which the TSI of the installed path/s is configured. Within this context three failure scenarios are considered and separately described: in the first scenario the failures occur simultaneously, in the second scenario the failures occur nearly simultaneously while in the third scenario the failures occur at different times.

[0027] Referring initially to Figs. 4, consider the case where two failures (SF1 and SF2) occur exactly at the same time instant. For simplicity, paths (a) and (b) before the occurrence of the failures, are allocated in a manner similar to what described for Fig. 1 and therefore the description of the allocations will not be repeated here. Upon the occurrence of the first failure (SF1) on the span 6-7, the node 6 (Fig. 4.1) will send a failure signaling (SF,7,6,L,ID) from the West side whereas, upon the second failure (SF2) on the span 7-8, node 8 will send a simultaneous failure signaling (SF,7,8,L,ID) from the East side (Fig. 4.2).

[0028] At the time when each of the two signaling with "Idle" code, which were generated by the switching nodes, is received by the termination nodes of the paths to be protected, squelching of the local termination (if any) of the LP channel corresponding to the HP channel allocated in the failed span to which the signaling is referred, takes place; while, at the nodes designed to realize the pass-through of the LP channels, the squelching of the local termination (if any) of the LP channel corresponding to the HP channel allocated in the failed span to which the signaling is referred takes place and the subsequent pass-through connection also takes place. The actions just described (squelching and squelching + pass-through) are removed both from the path termination nodes and from the pass-through nodes, as soon as such nodes receive the second signaling generated by the switching nodes.

[0029] When (Fig. 4.3) the signaling (SF,7,6,L,ID) containing the "Idle" code of SF1 reaches node 8, node 8 (Fig. 4.4) will send a signaling containing BR&SW (BS) Status Code of the type SF,7,8,L,BS. The same will be for node 6 (Fig. 4.5): as soon as it receives signaling (SF,7,8,L,ID) containing the "Idle" code of SF2, it will send a signaling containing BR&SW Status Code (BS) of the type SF,7,6,L,BS.

[0030] At the time when one of the two signalings with BS code generated by switching nodes is received by the termination nodes of the HP paths to be protected, the squelching of the local termination (if any) of the LP channel to be used for the protection, that was chosen according to one of the criteria described above, takes place; while, at the nodes designed to realize the pass-through of the LP channels, the squelching of the local termination (if any) of the same LP channel will take place and also the subsequent pass-through connection, will take place.

[0031] The Bridge & Switch action that is performed on the LP channel chosen according to the same criterion as above, is performed by every termination node of the HP paths to be protected, as soon as both signalings with BS Code (SF,7,6,L,BS and SF,7,8,L,BS) are detected at the two sides of the node itself.

[0032] Thus, a stable state of the protected ring has been achieved.

[0033] Referring initially to Figs. 5, consider the case where two failures (SF1 and SF2) occur nearly at the same time instant (or anyway failure SF2 occurs before the situation following SF1 is stabilized). For simplicity, paths (a) and (b) before the occurrence of the failures, are allocated similarly to what described for Fig. 1 and therefore the description of the allocations will not be repeated here. Upon the occurrence of the first failure (SF1) in span 6-7, the node 6 will send a failure signaling (SF,7,6,L,ID) from the West side and, similarly, it will send another failure signaling (SF,6,7,L,ID) from East side. See Figs. 5.1 and 5.2.

[0034] Suppose (Fig. 5.3) that the failure signaling (SF,6,7,L,ID) from the East side is able to reach node 8 before the second failure (SF2) occurs in span 7-8, which results in node 7 isolated. Upon the second failure (SF2), node 8 (node adjacent to the failure) will send a corresponding failure signaling (SF,7,8,L,ID) from its East side. Anyway, the signaling of the second failure will follow the first failure one (Fig. 5.4).

[0035] As soon as the signaling (SF,6,7,L,ID; SF, 7,6,L,ID) containing the "Idle" code of the first failure reach the termination node 2 (Fig. 5.5) of path (a), this node will perform the BR&SW action by utilizing the LP AU-4 corresponding to the span affected by the first failure (LP AU-4#7, in this instance). However, as soon as also the new signaling (SF,7,8,L,ID) of the second failure (SF2) reaches node 2, the BR&SW action, just realized, is removed (Fig. 5.6).

[0036] Analogously (Fig. 5.7), as soon as the signaling (SF,6,7,L,ID; SF,7,6,L,ID) containing the "Idle" code of the first failure reach the termination node 4 of the path (b), this node will perform the BR&SW action by utilizing the LP AU-4 corresponding to the span affected by the first failure (LP AU-4#6 in this case). However, as soon as also the new signaling (SF,7,8,L,ID) relating to the second failure (SF2) reaches node 4, the BR&SW action, just realized, is removed (Fig. 5.8).

[0037] Obviously, the actions to be taken before the just described temporary BR&SWs are the squelching of the local termination (if any) of the LP channel associated with the span 6-7 both on the termination nodes of the paths to be protected, and on the nodes designed to realize the pass-through, as well as the pass-through connection of the LP channel itself: in order a node to perform such actions, the reception of at least one of the

two signaling with "Idle" code generated by the switching nodes is enough to the interested node.

[0038] At the same time, when the signaling (SF,7,6,L, ID) containing the "Idle" code of the first failure reaches node 8, node 8 will send a signaling containing the BR&SW (BS) status code of the type SF,7,8,L,BS (Fig. 5.9). The same will be for node 6: as soon as it receives the signaling (SF,6,7,L,ID) containing the "Idle" code of the first failure, it will send a signaling containing the BR&SW (BS) Status Code of the type SF,7,6,L,BS (Fig. 5.10).

[0039] Because of the presence of the new signaling (SF,7,8,L,ID) concerning the second failure (SF2), node 6 will change again its signaling from SF,7,6,L,BS to SF, 7,6,L,ID (Fig. 5.11).

[0040] At this stage both the signaling transmitted by node 8 containing the BR&SW Status Code and the two consecutive signaling transmitted by node 6 respectively containing the BS&SW (BS) and the "Idle" status codes are present in the ring. The signaling containing the BS Status Code result, at the nodes detecting them, in the squelching of the local termination (if any) of the LP channel chosen for the protection according to one of the aforesaid criteria (for instance the LP channel corresponding to the allocation used in the span affected by the first failure, AU-4#6), as well as in the pass-through of such LP channel at the nodes designed to perform such a function. It is to be noted that, among the two signaling that consecutively emitted by node 6, the one containing "Idle" code does not remove the squelching and pass-through actions activated by the previous signaling (with BS code), since both refer to the same failed span: SF,7,6,L,BS and SF,7,6,L,ID.

[0041] The node 9 (Fig. 5.12), receiving a signaling with BS code (SF,7,6,L,BS and SF,7,8,L,BS) from both its W and E sides, will perform the BR&SW action by utilizing the LP allocation related to one of the failured spans, for instance the one affected by the first failure (AU-4#6). Node 8, that receives the signaling containing the BS code (SF,7,6,L,BS) previously sent to it by node 6, will realize the BR&SW action (Fig. 5.13) by utilizing the LP allocation related to one of the failured spans, for instance the one affected by the first failure (AU-4#7). Some of the possible selection criteria have been mentioned above.

[0042] Since the request that reaches both node 9 and node 8 with "Idle" code, is related to the failured span already indicated in the preceding request (SF, 7,6,L,BS), the BR&SW action is maintained (Fig. 5.14).

[0043] When the request related to the second failure (SF2) and containing the BS code reaches node 6, the APS signaling is updated with the BS code, namely node 6 will send, from side W, the signaling SF,7,6,L,BS (Fig. 5.15).

[0044] Node 4, as soon as it receives signaling with BS code from both sides, will realize the BR&SW action by utilizing the LP allocation related to one of the spans affected by a failure, for instance the one affected by the first failure (AU-4#6).

[0045] Lastly, the node 2, as soon as it receives, from both its W and E sides, a signaling with BS code (SF, 7,6,L,BS and SF,7,8,L,BS), will perform the BR&SW action (Fig. 5.16) by utilizing the LP allocation related to one of the spans affected by a failure, for instance the one affected by the first failure (AU-4#7).

[0046] Thus, a stable state in the protected ring is obtained.

[0047] As said above, the scenario of the actions taken by the various nodes is different in the case where the failures do not occur at the same time. In this connection, two different sub-scenarios should be distinguished. With reference to Figs 1 to 3 and 6, the actions and the consequences related to the first sub-scenario are schematically listed below starting from a situation free of faults.

[0048] The first failure (SF1) occurs. The node 6 sends SF,7,6,L,ID from the W side. The node 7 sends SF,6,7,L,ID from the side E (Fig. 2).

[0049] SF,7,6,L,ID and SF,6,7,L,ID reach the termination nodes of paths (a) and (b). The termination nodes perform the BR&SW action for each path to be protected by utilizing the corresponding LP channels of the span affected by SF1. Path (a) is allocated on LP AU-4#7. Path (b) is allocated on LP AU-4#6 (Fig. 3).

[0050] The nodes 6 and 7 adjacent to the failure SF1 send respective signaling with BS code (SF,7,6,L,BS and SF,6,7,L,BS) and a stable scenario of ring protected against SF1 is obtained (Fig. 3).

[0051] SF2 occurs on span 7-8: node 7 is isolated (Fig. 6.1). Node 8 sends SF,7,8,L,ID from the side E (Fig. 6.2).

[0052] The BR&SW action (both the BR&SW and the pass-through at the intermediate nodes) performed for path (a) is removed (Figs. 6.3 and 6.4). The BR&SW action (both the BS&SW and the pass-through at the intermediate nodes) performed for path (b) is removed (Figs. 6.5 and 6.6).

[0053] Node 6 receives the signaling SF,7,8,L,ID and sends SF,7,6,L,ID (Fig. 6.7).

[0054] The node 8 receives from node 6 the signaling SF,7,6,L,ID and sends SF,7,8,L,BS (Fig. 6.8).

[0055] Node 6 receives the signaling SF,7,8,L,BS and sends SF,7,6,L,BS (Fig. 6.9).

[0056] Nodes 2 and 8 receive the signaling SF,7,8,L, BS and SF7,6,L,BS and perform the BR&SW action by utilizing for instance the LP channels with AU-4 corresponding to that of the first failed span (LP AU-4#7). The scenario becomes stable for the path (a) (Figs. 6.10 and 6.11).

[0057] Nodes 4 and 9 receive the signaling SF,7,8,L, BS and SF,7,6,L,BS and perform the BR&SW action by utilizing for instance the LP channels with AU-4 corresponding to the one of the first failed span (LP AU-4#6). The scenario becomes stable for path (b) (Figs. 6.12, 6.13).

[0058] The squelching actions of the local termination

(if any) of the LP channel chosen for the protection according to one of the criteria already described and the subsequent pass-through of the same LP channel at the intermediate nodes come before the BR&SW actions just described and are performed with the rules already pointed out for the two previous scenarios.

[0059] With reference to Figs. 1 to 3 and 7, the actions and the consequences related to the second sub-scenario of double failure at different times will now be schematically listed in the following, still starting from a faultless situation.

[0060] The first failure (SF1) occurs. Node 6 sends SF,7,6,L,ID from the W side. Node 7 sends SF,6,7,L,ID from the E side (Fig. 1).

[0061] SF,7,6,L,ID and SF,6,7,L,ID reach the termination nodes of the paths (a) and (b). The termination nodes perform the BR&SW action for each path to be protected by utilizing the corresponding LP channels of the span affected by SF1. The path (a) is allocated on LP AU-4#7. The path (b) is allocated on LP AU-4#6 (Fig. 2).

[0062] The nodes 6 and 7 adjacent to failure SF1 receive the signaling with ID code (SF,7,6,L,ID and SF, 6,7,L,ID), send respective signaling with BS code (SF, 7,6,L,BS and SF,6,7,L,BS) and a stable scenario of ring protected against SF1 is obtained (Fig. 3).

[0063] SF2 occurs in span 7-8: node 7 is isolated (Fig. 7.1). Node 8 sends SF,7,8,L,ID from the E side (Fig. 7.2).

[0064] The node 8, as a node adjacent to the failure and as a termination node, evaluates whether the already protected paths can still be protected. In the affirmative, no action is taken; in the negative, the BR&SW action is removed (Fig. 7.3).

[0065] Node 9 receives the SF,7,8,L,ID request and evaluates whether the already protected paths can still be protected. In the affirmative, no action is taken; in the negative, the BR&SW action is removed (Fig. 7.4).

[0066] Node 2 receives the SF,7,8,L,ID request and evaluates whether the already protected paths can still be protected. In the affirmative, no action is taken; in the negative, the BR&SW action is removed (Fig. 7.5).

[0067] Node 4 receives the SF,7,8,L,ID request and evaluates whether the already protected paths can still be protected. If yes, no action is taken; if not, the BR&SW action is removed (Fig. 7.6).

[0068] As soon as node 6 receives the SF,7,8,L,ID signaling, it updates its request by inserting the "Idle" status code, of the type SF,7,6,L,ID. After a further signaling exchange, the nodes adjacent to the failure update the respective signaling by inserting "BR&SW" (BS) status code.

[0069] At this point only signaling with BS code are traveling in the ring that is failured by SF1 and SF2 and therefore a stable scenario for paths (a) and (b) towards the failures SF1 and SF2 has been achieved.

[0070] It will be recognized that the first sub-scenario results in a rather simple implementation since it is not necessary to store the failure "history" but, at the same time, traffic is not safeguarded in an optimal manner because the BR&SW is always removed. On the contrary, the second sub-scenario safeguards the traffic in a better manner but it is more difficult to be implemented because the traffic "history" shall be stored.

[0071] Having analyzed in detail the single-failure and the double-failure situations (simultaneous, nearly simultaneous or at different times), we will go on in describing schematically the actions that each network node must perform (and the corresponding consequences) when the failures are cleared and the ring functionality is restored.

[0072] Start from a stable situation of two failures SF1, SF2: in this situation, node 7 is isolated (Fig. 8.1) and only signalings with BS code (SF,7,8,L,BS and SF,7,6,L, BS) are traveling in the ring. Consider to clear first SF1: the node 7, no longer isolated, begins to send the APS signaling with "Idle" code related to the span affected by a failure (SF2) still present between the nodes 8 and 7 (SF,8,7,L,ID) (Fig. 8.2).

[0073] Since the LP allocation of the span 7-6 had been chosen, the BR&SW (and squelching of the any local termination of the LP channel utilized) action at node 4 must be removed. Similarly, as soon as also the SF,8,7,L,ID signaling reaches the other path termination nodes (2, 9, 8), the BR&SW and any local squelching action is removed also at such nodes 2, 9, 8 (Figs. 8.3 to 8.5). The removal of "BR&SW" at the termination nodes is accompanied by the removal of the pass-through (and of any local squelching) from the intermediate nodes that have performed the pass-through of the LP channel heretofore utilized for the protection. Since the signalings present at the intermediate nodes are related to the same span affected by a failure, such nodes can perform, if required, the pass-through of the LP channels, related to the current failure, to be utilized for the path protection.

[0074] The node 8, as a node adjacent to the failure SF2, receives SF,8,7,L,ID and changes the code of its signaling from SF,7,8,L,BS to SF,7,8,L,ID (Fig. 8.6). Such signaling with ID code gradually reaches all the termination nodes (9, 2, 4) showing them in this way that a single failure (SF2) is present. The termination nodes in turn will execute the BR&SW action (Figs. 8.8 to 8.10) by utilizing the LP channels that correspond to the failured span (for path (a) the LP AU-4#6 will be utilized, for path (b) the LP AU-4#7 will be utilized).

[0075] The nodes (7, 8) adjacent to the failure still present (SF2) will send corresponding signaling with BS code (SF,8,7,L,BS and SF,7,8,L,BS) and a single-failure stable condition will be achieved (Figs. 8.11, 8.12).

[0076] As soon as also SF2 is cleared, the ring will reach the faultless stable condition (Figs. 8.13, 8.14), with the progressive removal of the "Bridge" and "Switch" actions from all the path termination nodes and the consequent signalings with "No Request, Idle" code (NR,9,8,S,ID and NR,6,7,S,ID) by all the ring nodes, including nodes (7, 8) adjacent to the just cleared failure

(SF2).

[0077] Start now from a stable situation of two failures SF1, SF2 (Fig. 9.1): in this situation node 7 is isolated and only signalings with BS code (SF,7,8,L,BS and SF, 7,6,L,BS) are traveling in the ring. Consider to clear SF2 first: node 7 (Fig 9.2), no longer isolated, begins to send the APS signaling with "Idle" code (SF,6,7,L,ID) related to the span affected by failure (SF1) still present between nodes 6 and 7.

[0078] Since just the LP allocation of the span 7-6 had been chosen, the BR&SW action at node 8 can be maintained (Fig. 9.3). Similarly, the SF,6,7,L,ID signaling reaches the other path termination nodes (9, 2, 4) but the BR&SW action is maintained also at such nodes 9, 2, 4 (Figs. 9.4 to 9.6).

[0079] The same processing is carried out at intermediate nodes that perform the pass-through of the LP channels used for the protection: the pass-through is maintained.

[0080] Finally, also node 6 adjacent to the failure SF1 receives SF,6,7,L,ID and will send the corresponding signaling with ID code (SF,7,6,L,ID), reaching a stable scenario with "BS" signalings all over the ring.

[0081] As soon as also SF1 is cleared, the ring will reach the faultless stable condition, with the progressive removal of the "Bridge" and "Switch" actions from all the path termination nodes and the consequent signaling with "No Request, Idle" code (NR,5,6,S,ID and NR, 8,7,S,ID) issued by all the ring nodes, including nodes (6, 7) adjacent to the just cleared failure (SF2). See Figs. 9.7 and 9.8.

[0082] In view of the above detailed description, relating to some cases of single or double failure, the person skilled in the art can easily devise the actions that every node must perform in the event of a failure on other spans and/or in the case where more than two failures occur. Naturally, the present invention is applicable to all these cases and its scope covers all these cases and is limited only by the following claims.

[0083] As far as the practical realization is concerned, it will be understood that all the actions performed by every node or network element are the known Pass-Through, Bridge and Switch, squelching of any terminations of the Low-Priority channels involved in the protection and transmission of signaling, substantially of known type, actions. Therefore, the implementation of the present method does not require to change the physical structure of the existing network elements used in ring networks protected against possible failures. Any modifications must be carried out at level of consequent actions performed by the nodes affected by the protection mechanism, according to signalings already provided for and present in the standardized protocol and on the ground of ring map information, already provided for and processed, as well as traffic map that carries the allocation time-slot information, in every ring span of the single path that is installed.

[0084] Finally it is pointed out that, although the present invention has been described in detail with reference to SDH synchronous transmission, it applies, in similar manner, to other types of synchronous transmission, typically SONET. The fact that this type of signals has not been taken into account in the description shall not be interpreted as a limitation but merely as an example and in order to render the description clear. Therefore, for the purposes of the present description and of the annexed claims, the terminology used for SDH transmissions will include at least the corresponding SONET terminology and shall be read in this perspective.

Claims

1. Method of re-routing a path in a transoceanic MS shared protection ring network in the event of a failure on a span of said path, said ring network comprising network elements connected in a ring configuration by fiber spans, said fiber spans comprising high-priority (HP) channels and low-priority (LP) channels, said method comprising the step of performing a ring switch action by the MS shared protection mechanism, **characterized in that** a time slot interchange mechanism (TSI) is provided in said ring network and **in that** said method comprises the step of re-routing the path over the time slot of the low-priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the span affected by the failure.

2. Method according to claim 1, in which a further span of the path becomes affected by a failure, **characterized by** comprising the steps of: i) releasing the present re-routing that was performed because of the first failured span; ii) selecting one of the failured spans; and iii) re-routing the failured path over the time slot of the low-priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the failured span that has been selected.

3. Method according to claim 1, in which a further span becomes affected by a failure, **characterized by** comprising the step of maintaining the re-routing action, performed because of the first failured span, should the persistency of the re-routing information be supported by the network elements of the ring network.

4. Method according to claim 2, **characterized in that** the step of selecting one of the failured spans comprises the step of considering the two spans adjacent to the switching nodes that are able to communicate with the termination nodes of path to be protected in the case where at least one further span of the path becomes affected by a failure.

5. Method according to any of claims 2 to 4, **characterized in that** the step of selecting one of the failured spans comprises the step of selecting the failured span adjacent to the switching node having higher or lower node identification ID.

6. Method according to any of claims 2 to 4, **characterized in that** the step of selecting one of the failured spans comprises the step of selecting the failured span adjacent to the switching node that comes first or last in the network ring map.

7. Method according to any of claims 2 to 4, **characterized in that** the step of selecting one of the failured spans comprises the step of selecting the failured span adjacent to the far west (W) or far east (E) switching node in the ring network.

8. Network element of a transoceanic MS shared protection ring network, said ring network comprising other network elements connected each other in a ring configuration by fiber spans, said fiber spans comprising high-priority (HP) channels and low-priority (LP) channels, said network element comprising means for performing ring switch actions, namely pass-through, bridge or switch actions, upon reception of corresponding signalings and means for generating and sending proper signalings in response to reception of corresponding signalings, a path being installed in said ring network, **characterized in that** a time slot interchange mechanism (TSI) is provided in said ring network and **in that** said network element comprises means for, in case of failure in a span of the installed path, re-routing the path over the time slot of the low priority (LP) channels corresponding to the time slot of the high priority (HP) channels of the failured span.

9. Network element according to claim 8, in which a further span of the path becomes affected by a failure, **characterized by** comprising: i) means for releasing the re-routing action performed because of the first failured span; ii) means for selecting one of the failured spans; and iii) means for re-routing the path over the time slot of the low priority (LP) channels corresponding to the time slot of the high-priority (HP) channels of the failured span which has been selected.

10. Network element according to claim 8, in which a further span of the path becomes affected by a failure, **characterized by** comprising means for maintaining the re-routing action, performed because of the first failured span, should the persistency of the re-routing information be supported by the network elements of the ring network.

11. Network element according to claim 9, **characterized in that** said means for selecting one of the failured spans comprise means for considering the two spans adjacent to the switching nodes able to communicate with the termination nodes of path to be protected in the case where at least one further span of the path becomes affected by a failure.

12. Network element according to claim 9, said network element being a path termination node, **characterized by** comprising means for performing a Bridge&Switch action upon reception of two signalings comprising corresponding bridge requests with Bridge&Switch status code (BS) related to different spans.

13. Network element according to claim 9, said network element being a path non-termination node, **characterized by** comprising means for performing a pass-through action upon reception of at least one signaling comprising a bridge request with a Bridge&Switch status code (BS)

14. Network element according to claim 8 or 9, said network element being a path termination node, **characterized by** comprising means for performing a Bridge&Switch action upon reception of two signalings comprising corresponding bridge requests with Idle status code related to the same span.

15. Network element according to claim 9, said network element being a path non-termination node, **characterized by** comprising means for performing a pass-through action upon reception of at least one signaling comprising a bridge request with Idle status code.

16. Ring network comprising one or more network elements according to any of claims 8 to 15.

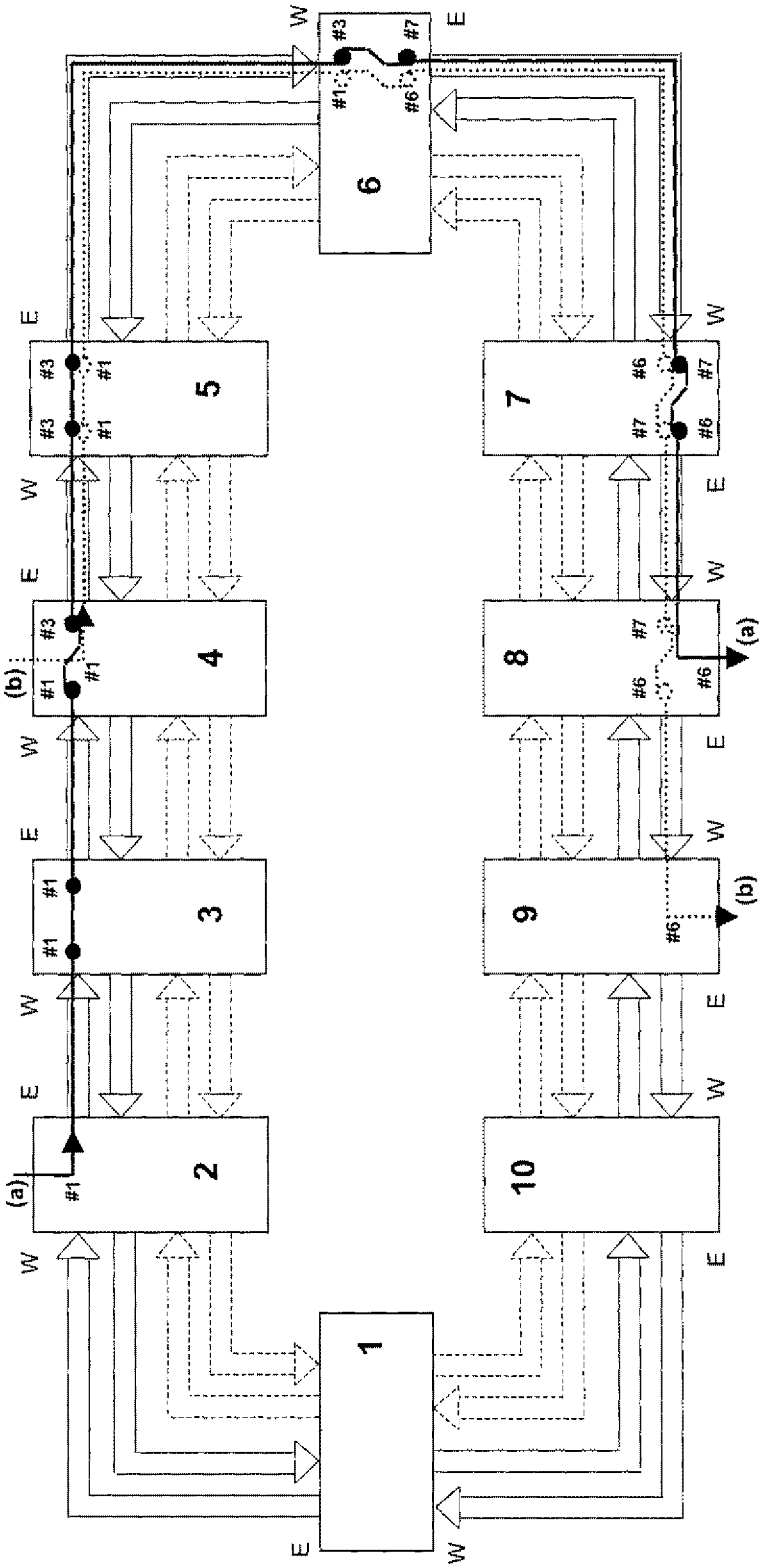


Fig. 1

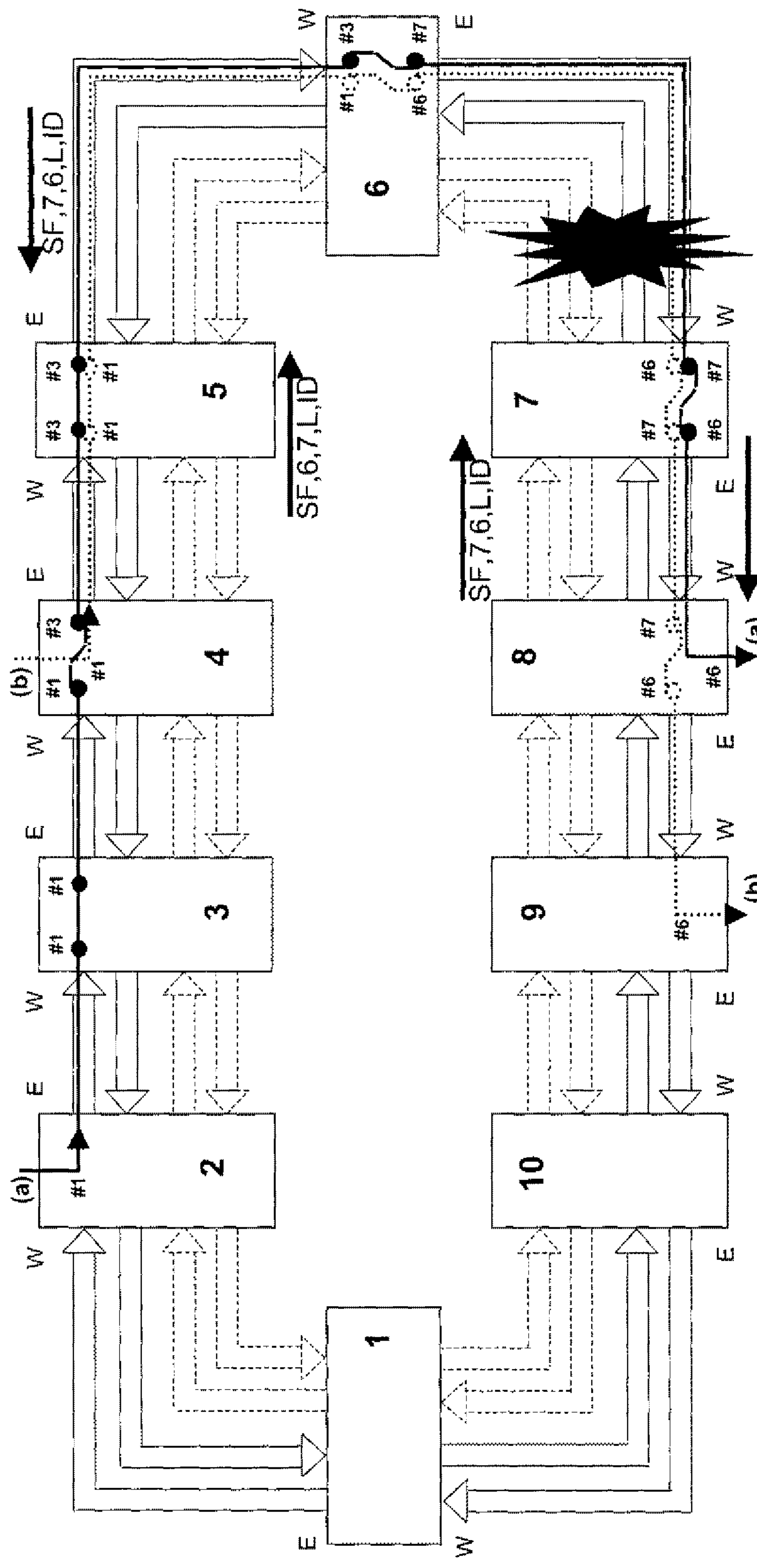


Fig. 2

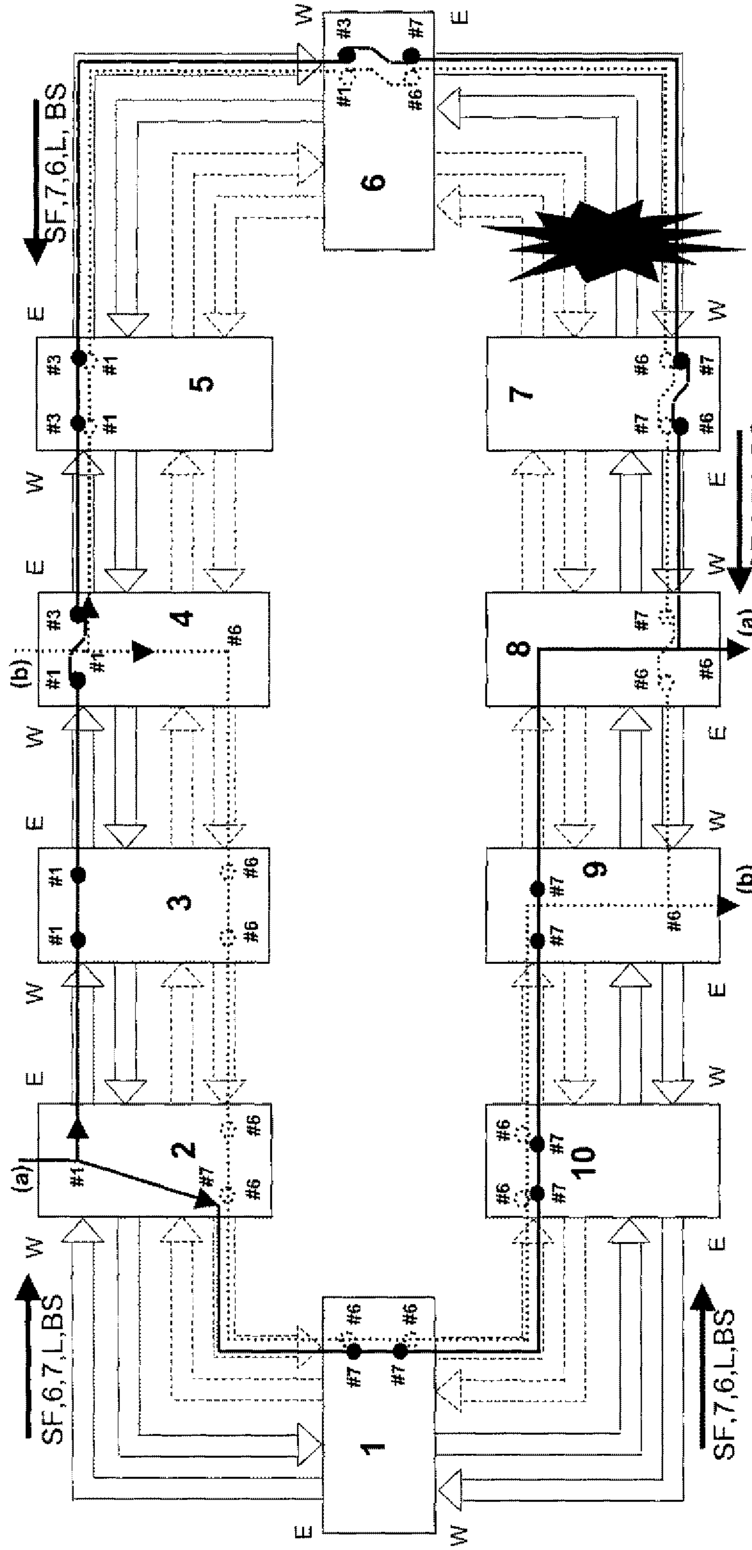


Fig. 3

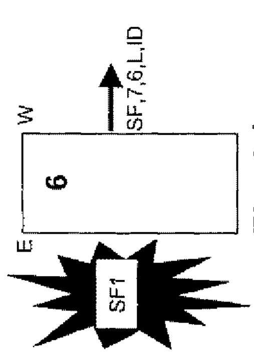


Fig. 4.1

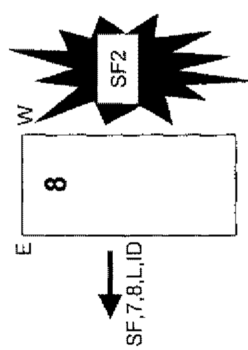


Fig. 4.2

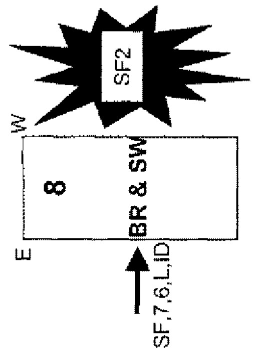


Fig. 4.3

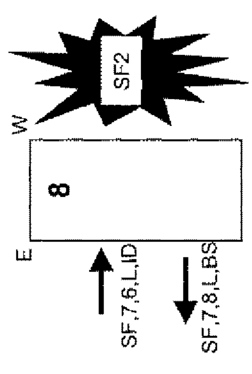


Fig. 4.4

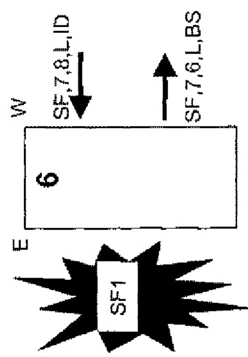


Fig. 4.5

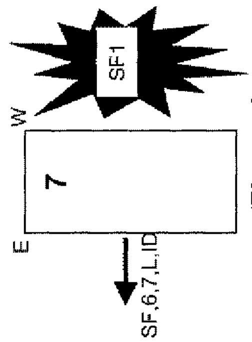


Fig. 5.1

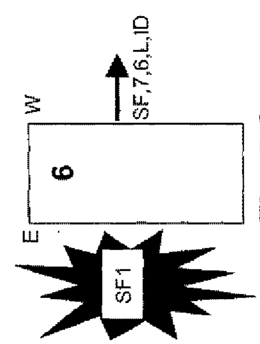


Fig. 5.2

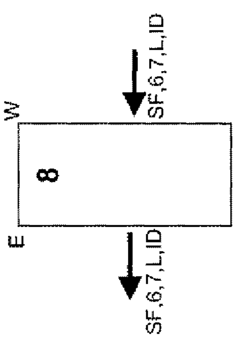


Fig. 5.3

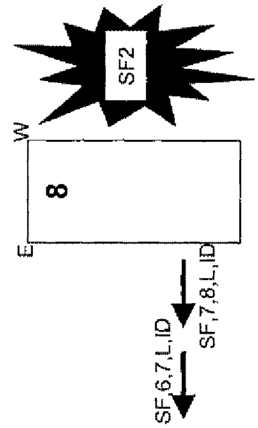


Fig. 5.4

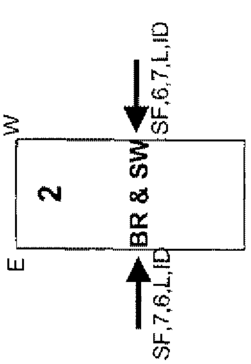


Fig. 5.5

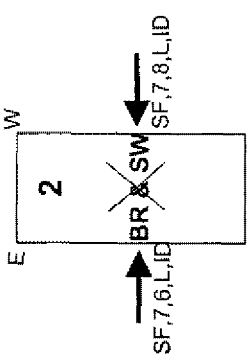


Fig. 5.6

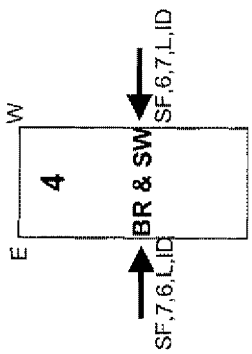


Fig. 5.7

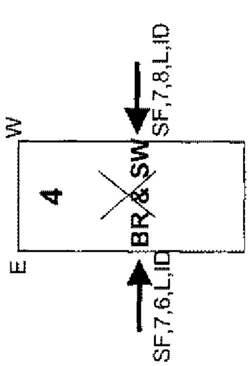


Fig. 5.8

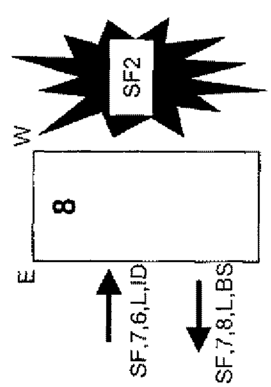


Fig. 5.9

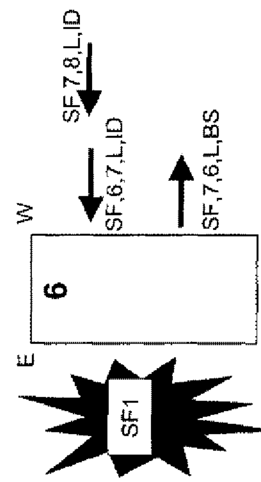


Fig. 5.10

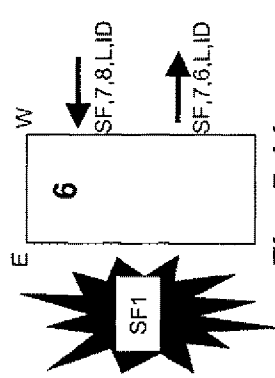


Fig. 5.11

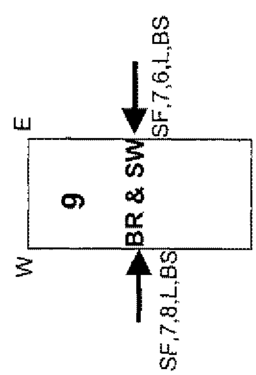


Fig. 5.12

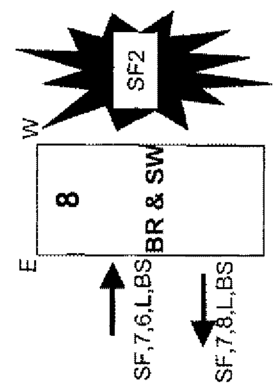


Fig. 5.13

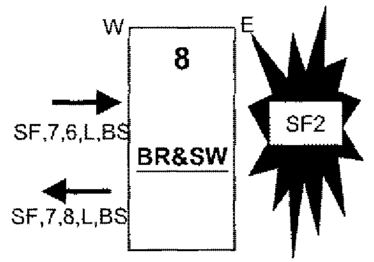


Fig. 5.14

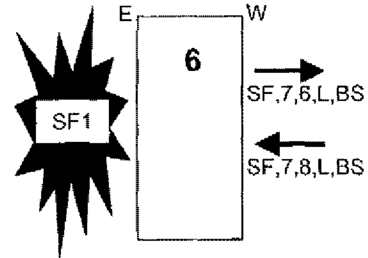


Fig. 5.15

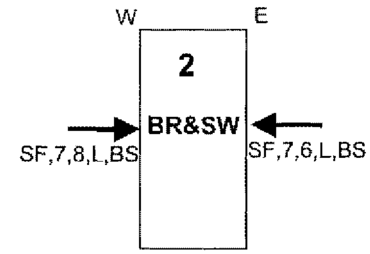


Fig. 5.16

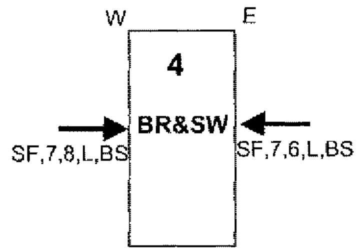


Fig. 5.17

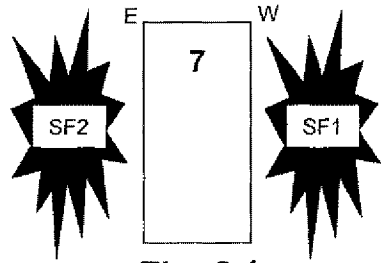
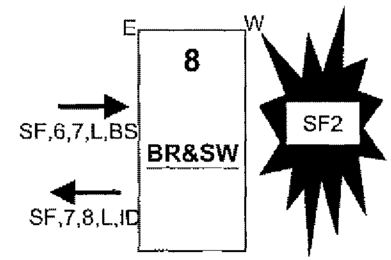


Fig. 6.1



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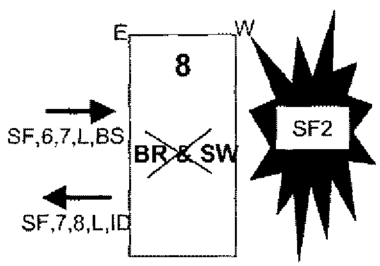

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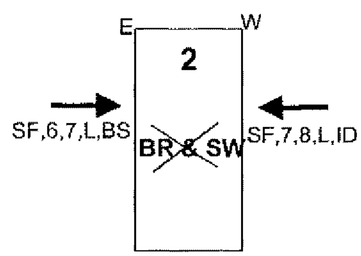

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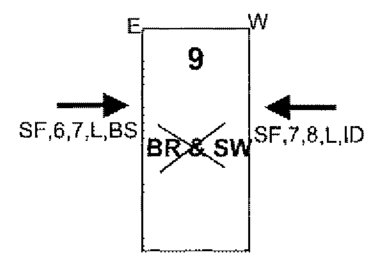

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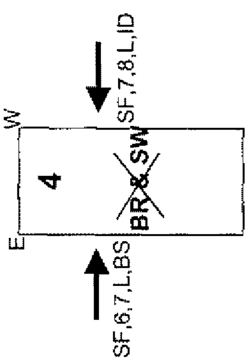

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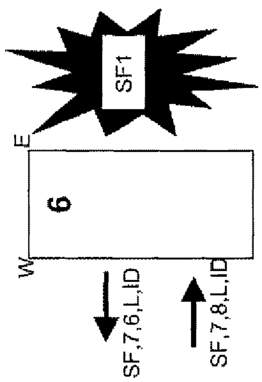

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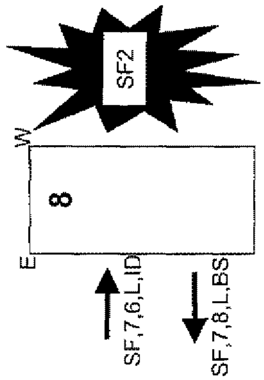

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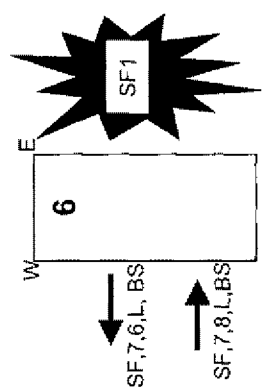

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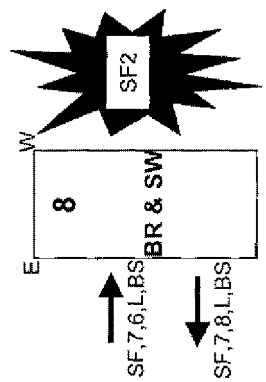

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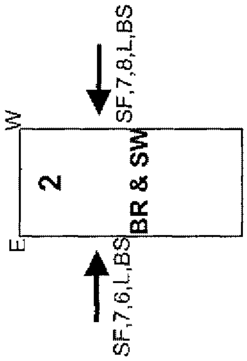

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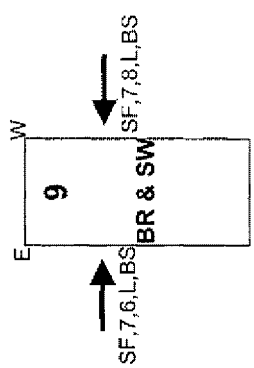

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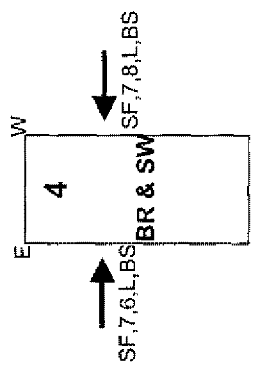

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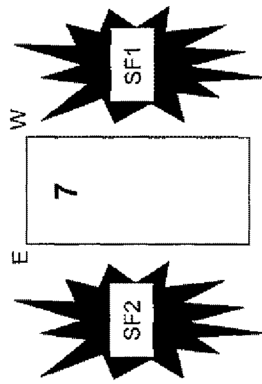

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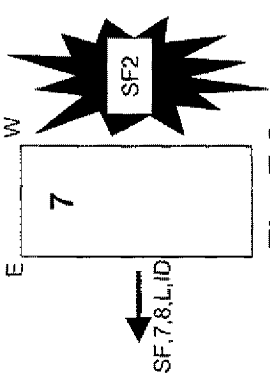

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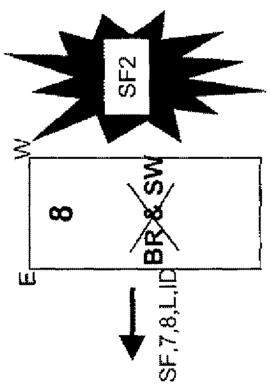

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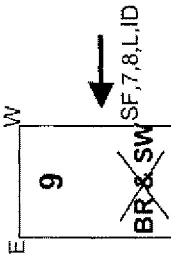

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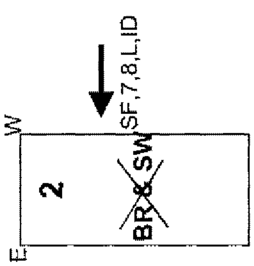

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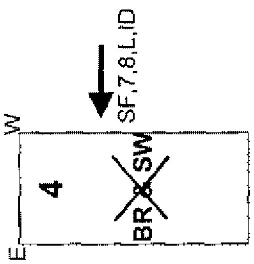

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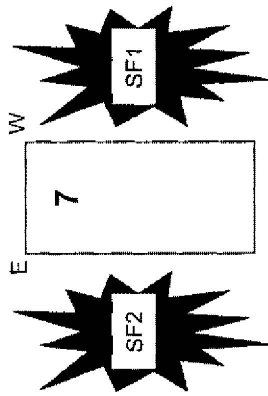

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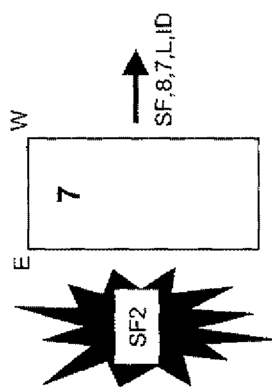

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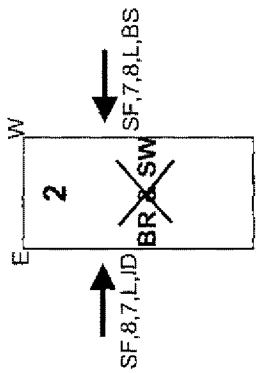

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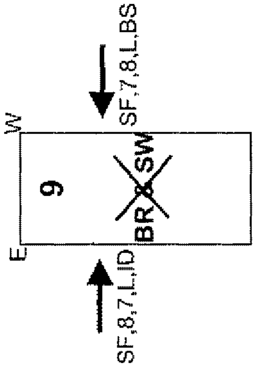

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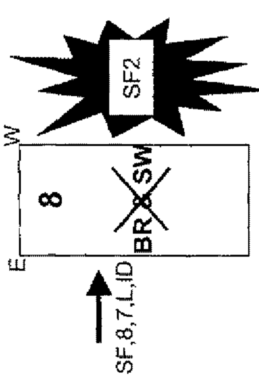

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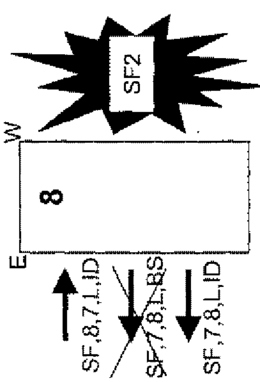

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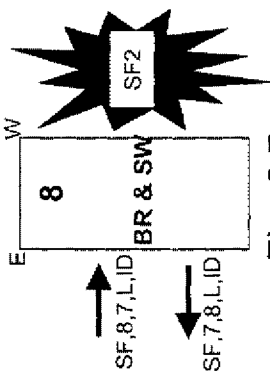

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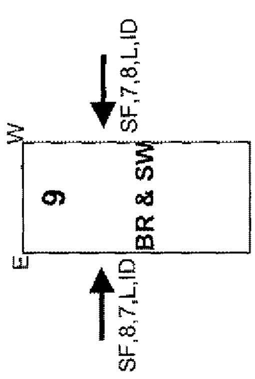

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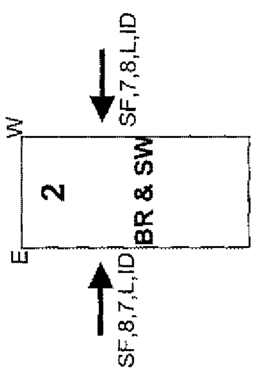

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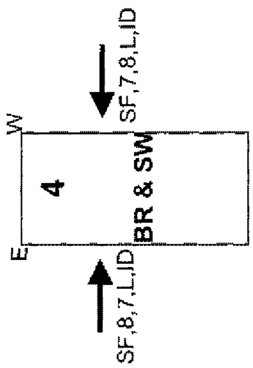

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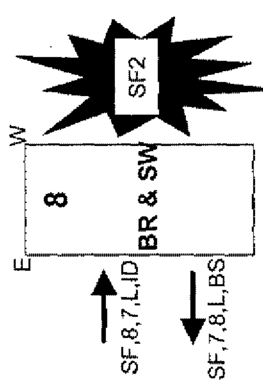

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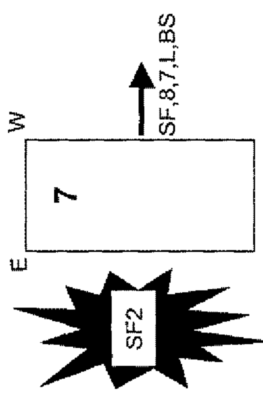

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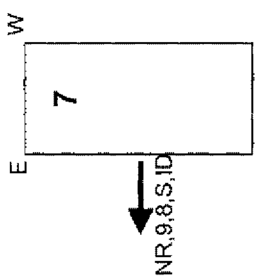

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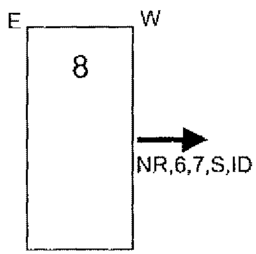

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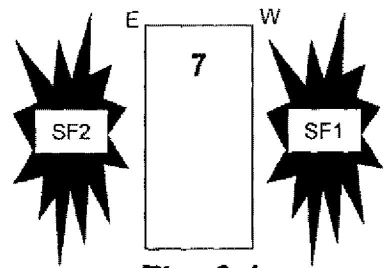

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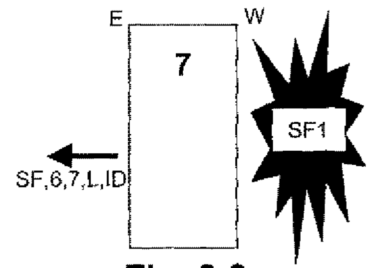

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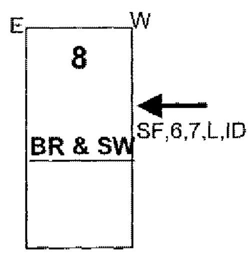

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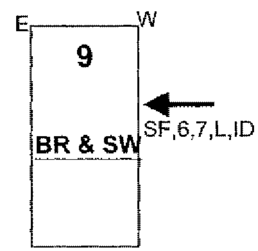

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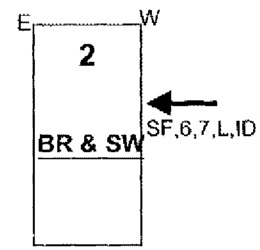

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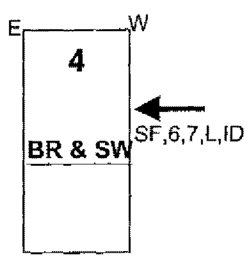

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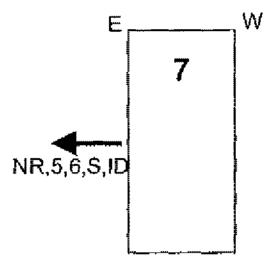

Fig. 9.7

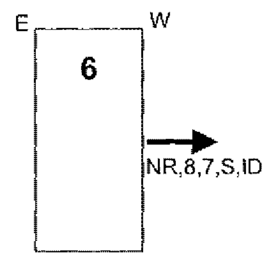

Fig. 9.8

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(54) Multicarrier system with adaptive bit-wise interleaving

(57) The present invention relates to a method for transmitting data streams of users via a transmission path in an OFDM system, whereby the data streams of respective users are transmitted in blocks, frequency hopping according to a predefined frequency hopping pattern for the respective transmission is performed, the block size for each user and within a hopping pattern can vary, and consecutive bits of the data stream to be transmitted are bit-wise interleaved, so that consecutive bits are transmitted in non-adjacent time slots and/or subcarriers within a block according to a predefined interleaving pattern. Thereby, the respective interleaving pattern is made adaptive to the respective frequency hopping pattern and/or the respective block size.

The present invention relates further to a transmitting station (1) for transmitting data streams of users and a receiving station (2) for receiving data streams of users for carrying out this method.

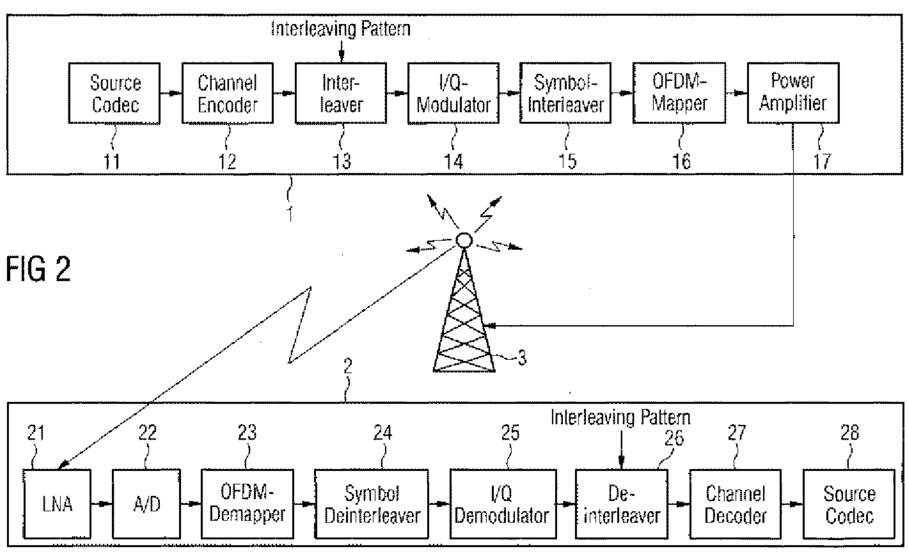

FIG 2

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Description

[0001] The present invention relates to a method for transmitting data in an orthogonal frequency division multiplex system (OFDM System). The present invention relates further to a transmitting station and a receiving station for carrying out this method.

[0002] The so-called orthogonal frequency division multiplex system (OFDM system) is widely used in broadcast systems like Digital Audio Broadcasting (DAB) and Digital Video Broadcasting (DVB) as e.g. described in EN 300 744 V1.2.1 of ETSI (European Telecommunication Standards Institute). This transmission system is also recommended for future wireless communication systems like BRAN (Broadband Radio Access Networks) and HIPERLAN (High Performance Radio Local Area Networks) as described in ETSI TS 101 475 V1.1.1 in order to provide high data rate services. Within these systems, the introduction of a bit-wise interleaver has increased the performance for high-level modulation schemes (e.g. 16QAM, 64QAM, Quadrature Amplitude Modulation), which are required for data transmission at a high-transfer rate.

[0003] A bit-wise interleaver interleaves consecutive and adjacent, respectively, bits of a data stream in a way, that adjacent bits of the data stream are transmitted in non-adjacent time slots and subcarriers, respectively. This has the advantage that associated transfer functions of respective consecutive bits (i.e. subcarrier and/or time slot number) are uncorrelated. In other words, particularly negative properties like e.g. deep fading of a specific transfer channel, do not take effect on consecutive bits, but rather only on single bits, e.g. of one transmitted symbol.

[0004] In wireless multi-user OFDM systems like e.g. BDMA systems (Band Division Multiple Access), for each user a predefined or fixed number of subcarriers and time slots can be assigned based on the required data rate. This assignment can change pseudo-randomly across the time/frequency-grid of an OFDM transmission path. This assignment is defined as one hopping pattern for one user, where at any one time slot some predefined or fixed subcarriers are allocated.

[0005] Figure 1 illustrates the transmission of data in an OFDM system with frequency hopping, whereby figure 1 shows a hopping pattern (assignment) for one user in a wireless multi-user OFDM system. Thereby, the time axis is divided into time slots of a predefined length and the frequency axis is divided into subcarriers of a predefined bandwidth.

[0006] As seen in figure 1, the data transmission for one user takes place in blocks, whereby each block has a length of a predefined number of time slots and a width of a predefined number of subcarriers. According to a hopping pattern, the (frequency) location, i.e. the frequencies occupied by a respective block within the transmitting path, of each block changes pseudo- randomly.

[0007] As shown in figure 1, the transmission of one user can also take place in more than one block within one period of time slots. This is illustrated in figure 1 as a hatched and a massive block, whereby both blocks belong to the data transmission of one user.

[0008] For each user, the respective corresponding hopping pattern can be repeated within any period of time, e.g. within one frame, one superframe or any other predefined fixed time period. In order to reduce the control burden, the hopping pattern for the respective user is assigned during a link initialisation and establishment phase and it does usually not change before the respective link (i.e. data transmission of one user) is released.

[0009] The block size, i.e. number of time slots and number of subcarriers, can be different for different users, it can also change from block to block (and between the different hopping steps, respectively) within a hopping pattern. These parameters are dependent of the required data rate and the resource management of the transmitting station (base station).

[0010] However, introducing a bit-wise interleaver in such an OFDM system, which performs frequency hopping as described in relation to figure 1, has the disadvantage, that, particularly at a relative small block size, the associated channel transfer functions can not be kept uncorrelated, since the time slot and/or subcarrier distance between the consecutive bits can not been made large enough.

[0011] Thus, due to the limited number of subcarriers at one specific time slot and the limited block size, respectively, it is difficult to implement the bit-wise interleaver according to the prior art in an OFDM frequency hopping system.

[0012] From WO 00/35102 an interleaving/de-interleaving device and a method for a communication system are known. A device for sequentially storing input bit symbols of a given interleaver size in a memory at an address and reading the stored bit symbols from the memory is provided. This known implementation method for an interleaver can be used for example based on CDMA-2000 specification or for other IMT-2000 communication systems. However, it cannot find application for the design of interleaver patterns for multi-user OFDM hopping systems.

[0013] From US 6,125,150 a transmission system using code design for transmission with periodic interleaving is known. Thereby an OFDM transmission system provides a high level of performance on a variety of frequency selective channels by using a code having the characteristics of maximum PPD and maximum PECL. The codes are designed to allow high SNR sub-channels to carry their full potential of information which is then used to compensate for information lost on low SNR sub-channels. According to this known technology error control coding, modulation and interleaver are combined together to obtain better distance characteristics, where some subcarriers may carry more information and another sub-carriers may carry less information

depending on the channel transfer functions.

[0014] It is therefore the object of the present invention to provide a technique for transmitting data streams in an orthogonal frequency division multiplex system (OFDM system), whereby the performance of interleaving and therefore the performance of the transmission is improved.

[0015] The above object is achieved by a method of transmitting data streams of users via a transmission path in a OFDM system according to claim 1.

[0016] This object is further achieved by a transmitting station and a receiving station for carrying out this method according to claims 7 and 8.

[0017] The method for transmitting data streams of users via a transmission path in a OFDM system according to the present invention performs a data transmission. The time axis of the transmission path is divided into time slots The frequency axis of the transmission path is divided into subcarriers. The resource of the transmission path is used by a plurality of users. The data streams of the respective users are transmitted in blocks with a block size of a predefined length of time slots and a predefined number of subcarriers. Frequency hopping according to a predefined frequency hopping pattern for the respective transmission is performed. The frequency hopping pattern for a respective transmission can differ between different users and it also can differ between different times for the same user.

[0018] Further, the frequencies occupied by a respective block within the transmission path vary according to the frequency hopping pattern. The block size for each user within a hopping pattern can also vary. Consecutive bits of the data stream to be transmitted can be bit-wise interleaved such that consecutive bits are transmitted in non-adjacent time slots and/or subcarriers according to a predefined interleaving pattern.

[0019] According to the present invention, the respective interleaving pattern is thereby made adaptive (and can be a function of) to the respective frequency hopping pattern and/or the respective block size.

[0020] The transmitting station according to the present invention for transmitting data streams of users comprises an interleaving means for bit-wise interleaving consecutive bits of data streams according to a predefined interleaving pattern, whereby the interleaving means uses an interleaving pattern, which is made adaptive to the respective frequency hopping pattern and/or the respective block size.

[0021] The receiving station for receiving data streams of users, which are transmitted according to the above-mentioned method according to the present invention comprises an deinterleaving means for deinterleaving the received data streams into the original bit sequence according to a predefined interleaving pattern, whereby the deinterleaving means uses an interleaving pattern, which is made adaptive to the respective frequency hopping pattern and/or the respective block size.

[0022] The present invention has the advantage that the performance of high level modulation schemes, like 16QAM, 64QAM or higher, can be improved, since the bit-wise interleaving is made adaptive to the respective data transmission, i.e. the respective hopping pattern and/or block size of the respective transmission. Parasitic characteristics of respective channels are minimised, since e.g. deep fading ideally take only effect on single bits e.g. of one transmitted symbol.

[0023] Thereby, consecutive bits are transmitted within the same block, when the transmitted blocks are large enough, that means, when a block size allows to transmit consecutive bits in the same block so that the associated channel transfer functions keep uncorrelated.

[0024] Consecutive bits can also be transmitted in different blocks according to the interleaving pattern. Advantageously, consecutive bits are transmitted in different blocks, when the block size is very small, so that the associated channel transfer functions for consecutive bits can be kept uncorrelated.

[0025] Further advantageously, the interleaving pattern is made adaptive to the number of time slots of the respective block and/or the interleaving pattern is made adaptive to the number of subcarriers of the respective block.

[0026] In the following description a preferred embodiment of the present invention is explained in more detail in relation to the enclosed drawings, in which

Figure 1 shows an example of data transmission in an OFDM system with frequency hopping,

Figure 2 shows a block diagram of a wireless OFDM system according to the present invention,

Figure 3 shows a diagram of 16QAM and 64QAM mappings and the corresponding bit pattern,

Figure 4 shows one example of the mapping of one symbol into the time-frequency grid of OFDM, and

Figure 5 shows another example of the mapping of one symbol into the time-frequency grid of OFDM.

[0027] Figure 2 shows a schematic diagram of a wireless OFDM system according to the present invention, whereby a block diagram of a transmitting station 1 and a block diagram of a receiving station 2 are depicted.

[0028] The transmitting station 1 according to the present invention comprises a source codec 11 for coding the signals which have to be transmitted (e.g. audio or video signal) into a data stream of a digital signal, and a channel encoder 12 for encoding a data stream e.g. into a frame structure, adding redundancy bits, etc.

[0029] Then, the data stream is adaptively bit-wise interleaved by the interleaver 13 according to the present invention. The pattern for bit-wise interleaving the data stream is thereby made adaptive to predefined param-

eters of the respective transmission like frequency hopping pattern and block size. In other words, the pattern is a function of said parameters. The adaptive interleaving according to the present invention is described later in more detail with reference to Figs. 4 and 5.

[0030] After interleaving the data stream is modulated into symbols, e.g. according to the known I/Q modulation (In-phase/Quadrature modulation), by an I/Q modulator 14 and map into a time/frequency-grid by an OFDM mapper 16. Optionally, the data stream can be symbol-wise interleaved by a symbol interleaver 15 as known from the prior art in order to further improve the transmission performance.

[0031] The OFDM mapper 16 maps the modulated data stream into the time/frequency-grid according to the OFDM transmission system. Further, the OFDM mapper 16 determines the block size and the used frequency hopping pattern, both dependent e.g. upon a given users data rate and resource management in the transmitting station 1.

[0032] Then, the mapped data stream is than amplified by a power amplifier 17 and transmitted via a radio tower 3 over an air-interface to one or plurality of receiving stations.

[0033] The receiving and demodulating of data by the receiving station 2 is carried out in the inverse sequence.

[0034] Thereby, the signal transmitted by the radio tower 3 is received by an antenna comprising a low noise amplifier 21 (LNA). The received signal (comprising the data stream) is analogue/digital converted by an A/D converter 22.

[0035] Complementary to the OFDM mapper 16 of the transmitting station 1, the received signal is demapped by an OFDM demapper 23. The signal is thereby demapped according to the same pattern for mapping the data stream by the OFDM mapper 16 in order to reconstruct the original data stream.

[0036] If the signal is symbol-wise interleaved by the transmitting station 1, the signal has to be symbol-wise deinterleaved by a symbol deinterleaver 24.

[0037] After I/Q demodulation of the demapped data stream by the I/Q demodulator 25, the data stream is bit-wise deinterleaved by the deinterleaver 26. Thereby, the pattern for deinterleaving the data stream is made adaptive to the hopping pattern for mapping/demapping the signal; the pattern for bit-wise deinterleaving is similar to the interleaving pattern used by the transmitter 1 in order to get the original data stream.

[0038] Channel decoding and source decoding is performed by a channel decoder 27 and source decoder 28 similar to the source coding and channel coding of the transmitting station 1.

[0039] Figure 3 shows the principles of QAM (Quadrature Amplitude Modulation) based on the example of 16QAM and 64QAM.

[0040] For QAM the information is transmitted with an in-phase and a quadrature component Q. Thus, the carrier comprises respective to the information, which are transmitted, an in-phase (I) and an quadrature (Q) component. Thereby, dependent on the modulation scheme (e.g. 16QAM or 64QAM), one transmitted symbol carries 4 bits at 16QAM (respectively 2 bits for I- and Q-channel) and 6 bit at 64QAM (respectively 3 bits for I- and Q-channel) pursuant to the scheme as shown in the respective coordinate system of 16QAM and 64QAM; the bit order is termed by I1, Q1, I2, Q2 and I1, Q1, I2, Q2, I3, Q3, respectively.

[0041] Thereby, the high priority bits are I1 and Q1. For 16QAM, the low priority bits are I2 and Q2, for 64QAM, the low priority bits are I3 and Q3. At the example of 64QAM (encircled symbol 000011) it is illustrated, that the high priority bits are less susceptible against interferences than the low priority bits. If, e.g., this symbol is interfered, it could be decoded wrongly as an adjacent symbol, e.g. as 000010, 000111, 001011 or 000001. It is seen, that the high priority bits are always the same, namely 00. Thus, the high priority bis are more protected against interferences than the low priority bits.

[0042] Figure 4 shows one example of a pattern for mapping a data stream into a time/frequency-grid by the OFDM mapper 16 shown in figure 2.

[0043] In this example, the hopping pattern for one user is shown. Thereby, each use is assigned two blocks (shown as hatched and solid blocks). Since the respective blocks are very small, i.e. low number of subcarriers in this example, consecutive bits of one symbol, e.g. I1, Q1, I2, Q2 are transmitted in different blocks, interleaved according to a bit-wise interleaving pattern A.

[0044] Figure 5 shows a different hopping pattern for mapping a data stream into the time/frequency-grid.

[0045] Thereby, the block size differs between the single hopping steps. E.g. in the first block and the third block are transmitted two consecutive bits (respectively I1, Q1 and I2, Q2), since the block size is large enough. In this case, bit-wise the interleaving happens according to an interleaving pattern B.

[0046] Thus, the design rule of a bit-wise interleaver 13 as shown in figure 1 is as follows:

- Adjacent coded bits from channel encoder are mapped onto non-adjacent subcarriers or non-adjacent time slots. The frequency separation (distance) of the chosen subcarriers or the time separation of the chosen time slot has to be far enough in order to keep the associated channel transfer functions uncorrelated.

- Adjacent coded bits from channel encoder are mapped alternatively on high or low priority bits. By this way, long runs of low reliability bits are avoided.

- The coded bits are placed at all available subcarriers and time slots on OFDM time/frequency-grid within the depth, i.e. the time/frequency/block dis-

tance between consecutive bits, of the interleaver are used.

- The bit-wise interleaver pattern is made adaptive to the hopping pattern in order to achieve better system performance.

[0047] The present invention has the advantage, that the performance of high level modulation schemes, like 16QAM, 64QAM or higher, can be improved, since the bit-wise interleaving is made adaptive to the respective data transmission, i.e. the respective hopping pattern and/or block size of the respective transmission. Parasitic characteristics of respective channels are minimised, since e.g. deep fading ideally take only effect on single bits e.g. of one transmitted symbol.

[0048] According to invention therefore a new design rule is proposed for a bit-wise interleaver for multi-user OFDM hopping systems. Instead of placing data bits belonging to one I/Q symbol or adjacent symbols on different sub-carriers at the same timeslot, data bits belonging to one I/Q symbol or adjacent symbols can be placed at different timeslots or at different blocks within the bit-wise interleaver.

[0049] Furthermore, in multi-user OFDM hopping systems, where each user can be assigned different hopping patterns (depending upon a given user's data rate and resource management in the base station) at different times, it is proposed that the practical bit-wise interleaver pattern to be used for each user is variable and depends upon its assigned hopping pattern. In this way the optimal performance can be obtained.

[0050] Each sub-carrier thereby carries the same size of information. By using a bit-wise interleaver, the bits belonging to one symbol are interleaved. Therefore, the good performance can be achieved for error control coding. Therefore, the use of a bit-wise interleaver can be enabled for multi-user OFDM hopping systems.

Claims

1. Method for transmitting data streams of users via a transmission path in an OFDM system, whereby
the time axis of the transmission path is divided into time slots,
the frequency axis of the transmission path is divided into subcarriers,
the transmission path is used by a plurality of users,
the data streams of the respective users are transmitted in blocks with a block size of a predefined length of time slots and a predefined number of subcarriers,
frequency hopping according to a predefined frequency hopping pattern for the respective transmission is performed, whereby the frequency hopping pattern for respective transmission can differ between different users and differ between different times for the same user,
the frequencies occupied by a respective block within the transmission path vary according to the frequency hopping pattern,
the block size for each user and within a hopping pattern can vary, and
consecutive bits of the data stream to be transmitted are bit-wise interleaved, so that consecutive bits are transmitted in non-adjacent time slots and/or subcarriers according to a predefined interleaving pattern,
characterised in
that the respective interleaving pattern is made adaptive to the respective frequency hopping pattern and/or the respective block size.

2. Method according to claim 1,
characterised in
that consecutive bits are transmitted within the same block, in case the block size is large.

3. Method according to claim 1,
characterised in
that consecutive bits are transmitted in different blocks according to the interleaving pattern.

4. Method according to claim 3,
characterised in
that consecutive bits are transmitted in different blocks, in case the block size is small.

5. Method according to one of the claims 1 to 4,
characterised in
that the interleaving pattern is made adaptive to the number of time slots of the respective block.

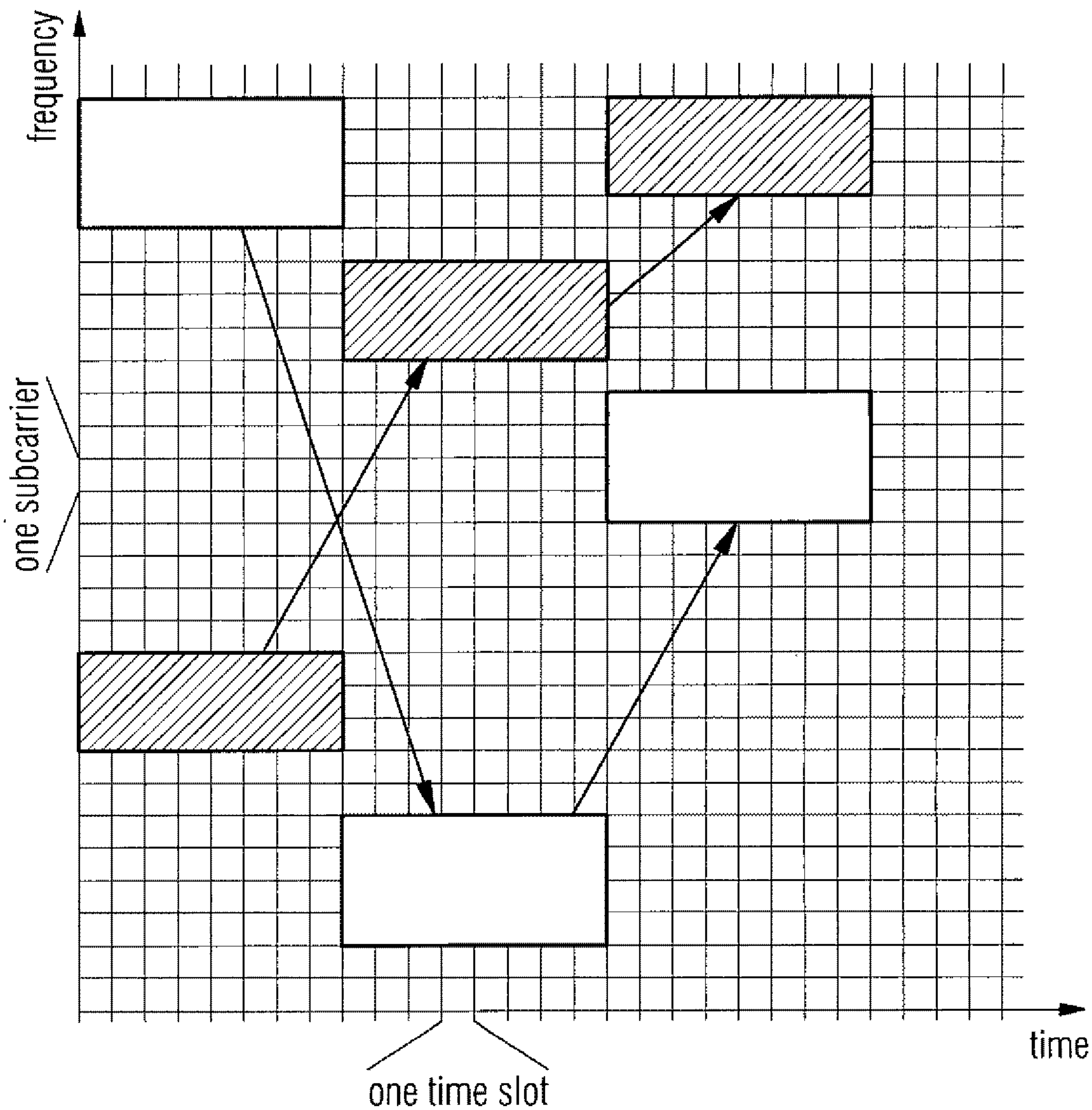
6. Method according to one of the claims 1 to 5,
characterised in
that interleaving pattern is made adaptive to the number of subcarriers of the respective block.

7. Transmitting station (1) for transmitting data streams of users by using a method according to anyone of the claims 1 to 6,
characterised by
an interleaving means (13) for bit-wise interleaving consecutive bits of data streams according to a predefined interleaving pattern, whereby the interleaving means (13) uses an interleaving pattern, which is made adaptive to the respective frequency hopping pattern and/or the respective block size.

8. Receiving station (2) for receiving data streams of users, which are transmitted using a method for transmitting data streams according to anyone of the claims 1 to 6,
characterised by
an deinterleaving means for deinterleaving the re-

ceived data streams into the original bit sequence according to a predefined interleaving pattern, whereby the deinterleaving means uses an interleaving pattern, which is made adaptive to the respective frequency hopping pattern and/or the respective block size.

FIG 1



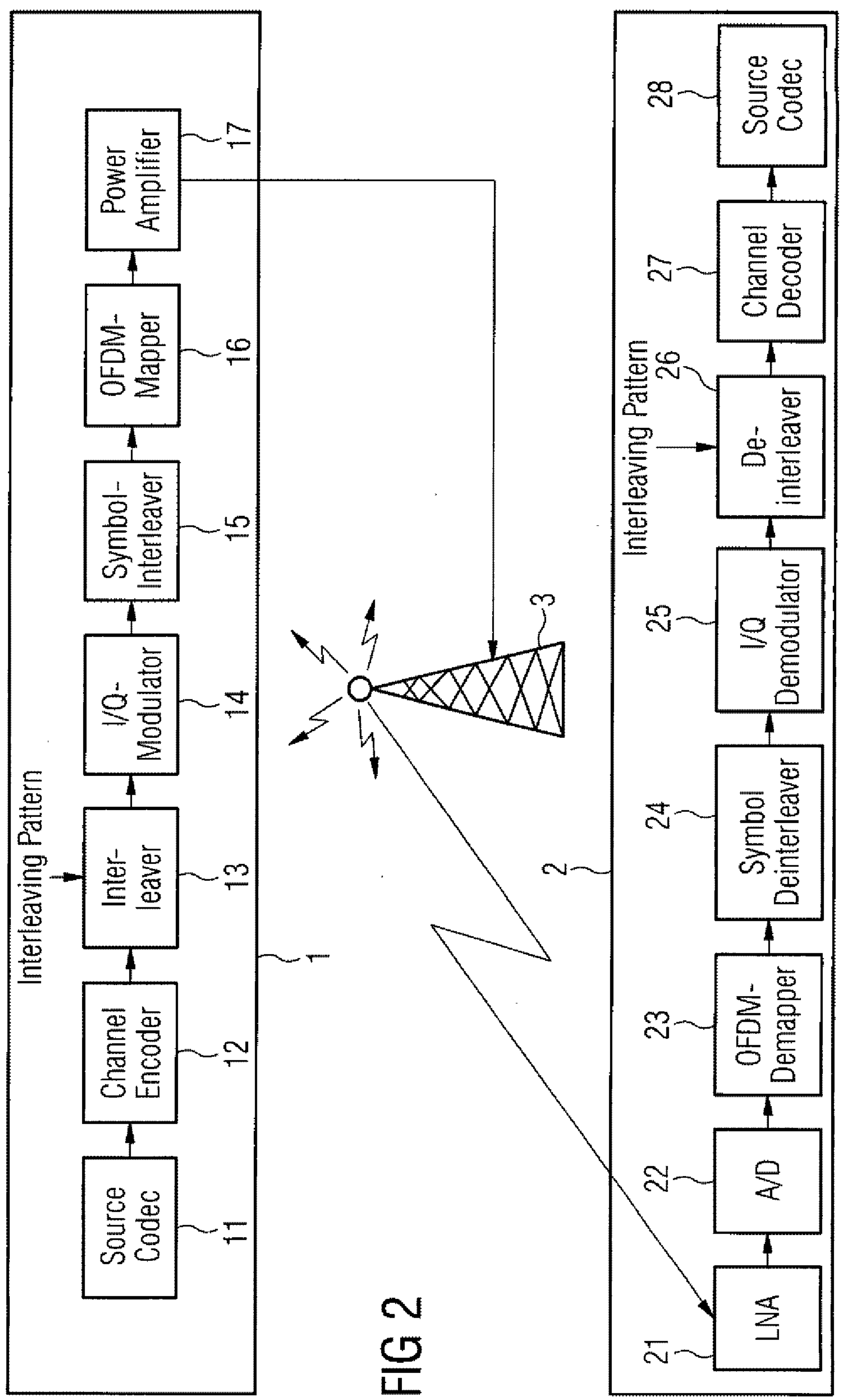

Interleaving Pattern
Source Codec
Channel Encoder
Inter-leaver
I/Q-Modulator
Symbol-Interleaver
OFDM-Mapper
Power Amplifier
11
12
13
14
15
16
17
1
3
2
Interleaving Pattern
LNA
A/D
OFDM-Demapper
Symbol Deinterleaver
I/Q Demodulator
De-interleaver
Channel Decoder
Source Codec
21
22
23
24
25
26
27
28

FIG 2

FIG 3

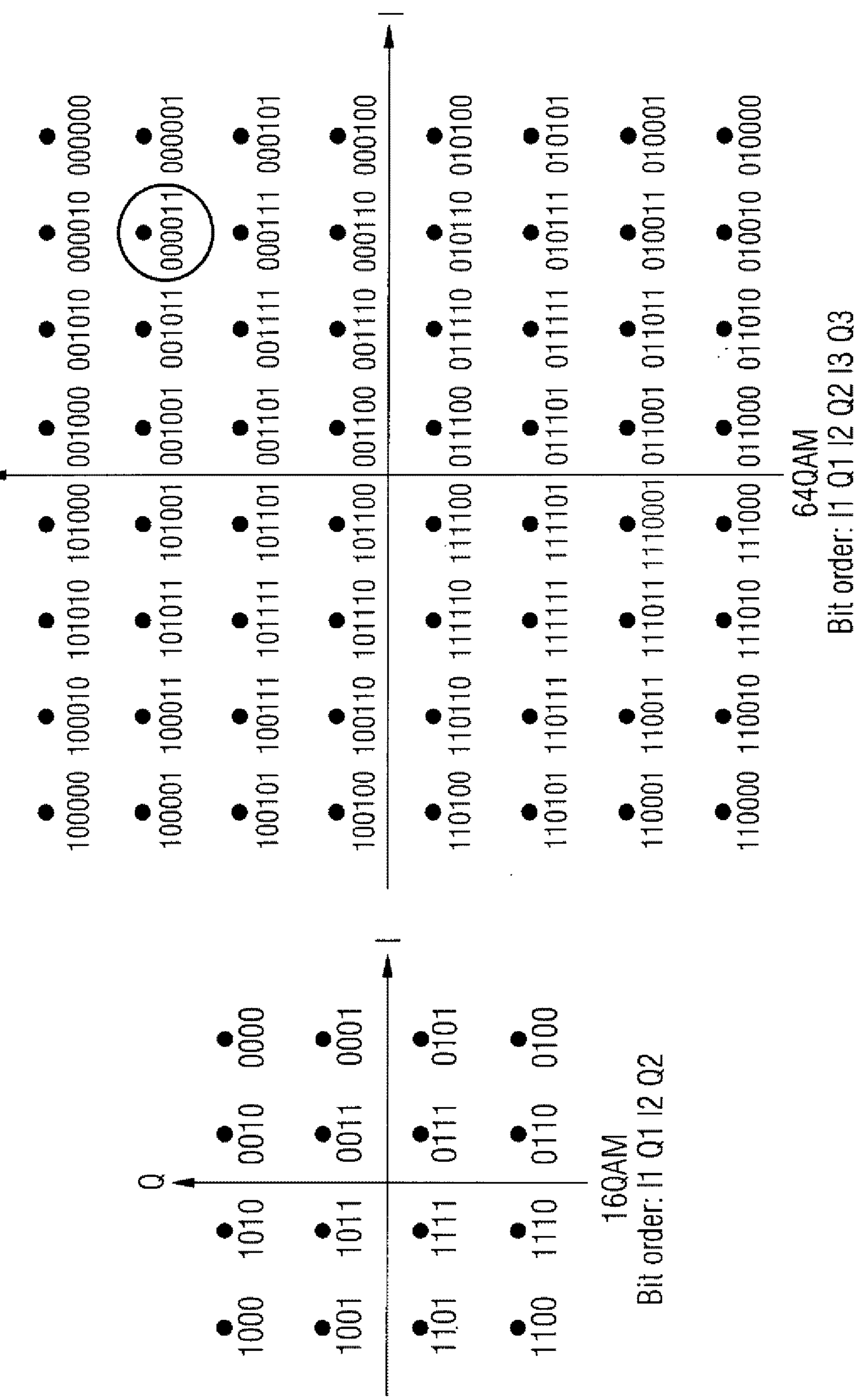


FIG 4

Interleaving Pattern A

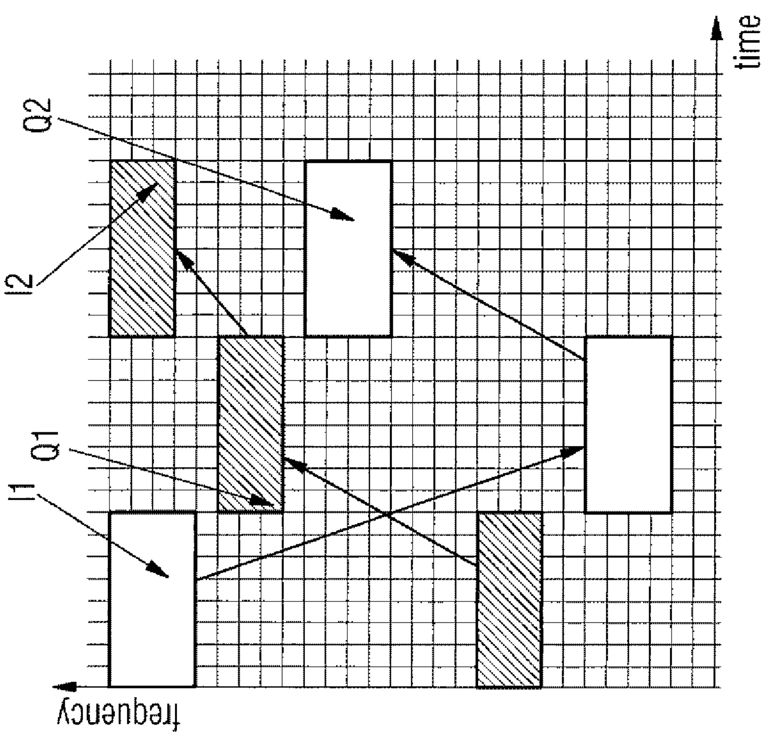
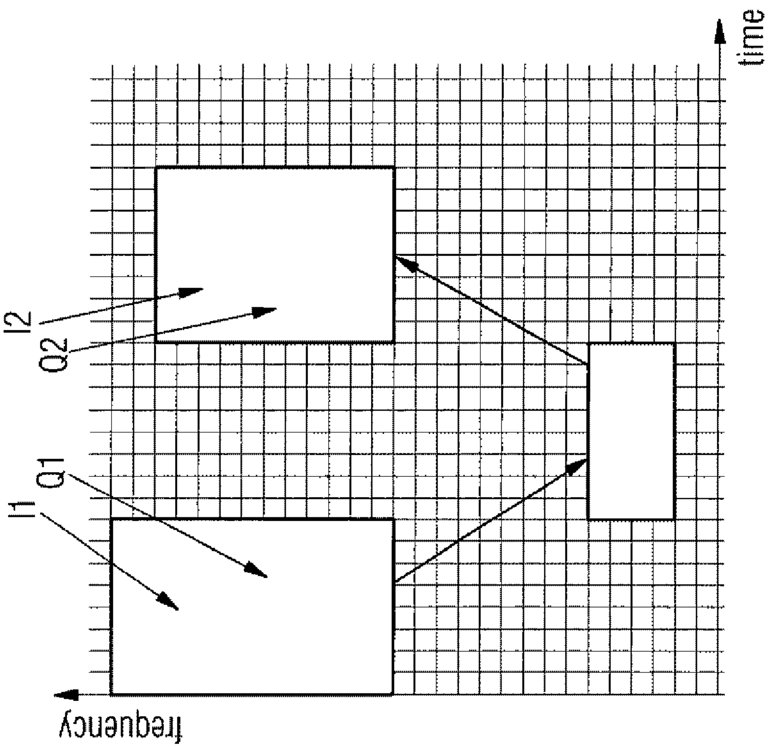


FIG 5

Interleaving Pattern B



European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 01 11 4041

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Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
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The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	17 October 2001	Papantoniou, A

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
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17-10-2001

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EPO FORM P0459

For more details about this annex : see Official Journal of the European Patent Office, No. 12/82

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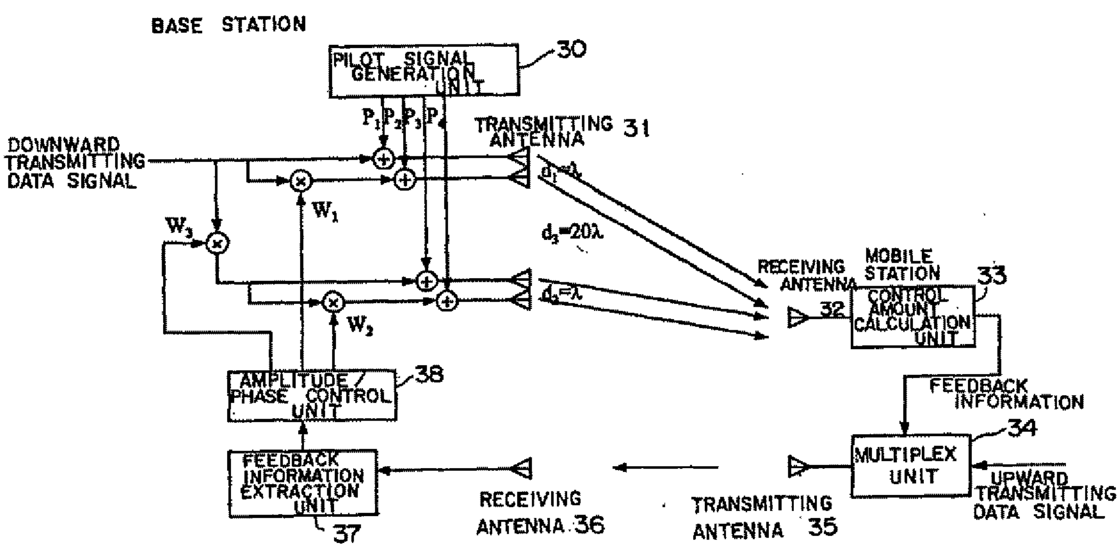
(54) **TRANSMISSION DIVERSITY COMMUNICATION DEVICE**

(57) The plurality of antennas of a base station used for transmitting diversity are divided into groups. Each antenna is located so that signals transmitted from antennas in the same group have a high fading correlation. Each antenna group is spaced so that a fading correlation between the groups may become low. Since signals transmitted from an antennas in the same group have high fading correlation, such signals suffer little from fading fluctuations and a low control speed is acceptable. However, control between the groups must be exercised at a high speed. Therefore, a mobile station that receives the signals of the base station feeds back feedback information for controlling fading fluctuations between the groups and information within a group to the base station at a high transfer rate and at a low transfer rate, respectively.



F I G. 4

EP 1 315 311 A1

Description

Technical field

[0001] The present invention relates to a transmitting diversity communications apparatus.

Background Art

[0002] Transmitting diversity in W-CDMA, which is the third-generation mobile communications system, adopts a method using two transmitting antennas.

[0003] Fig. 1 shows an example configuration of a transmitting diversity system using two transmitting antennas.

[0004] Mutually orthogonal pilot patterns P_1 and P_2 are transmitted from two transmitting antennas 1 and 2, respectively, as pilot signals, and channel impulse response vectors $\underline{h}_1$ and $\underline{h}_2$ from each antenna of a base station up to the receiving antenna of a mobile station are estimated by correlating each known pilot pattern to an incoming pilot on the receiving side of the mobile station.

[0005] A control amount calculation unit 10 calculates and quantizes the amplitude/phase control vector (weight vector) $\underline{w}=[w_1, w_2]$ of each transmitting antenna of the base station that maximizes power P expressed by the following equation (1) using these channel estimation values. Then, a multiplex unit 11 multiplexes the quantized weight vectors with an uplink channel signal as feedback information and transmits the signal to the base station. However, since there is no need to transmit both values w_1 and w_2, it is acceptable to transmit only value w_2 obtained by assigning w_1=1.

$$P = \underline{w}^H H^H H \underline{w} \tag{1}$$

$$H = [\underline{h}_1, \underline{h}_2] \tag{2}$$

[0006] In equation (2), $\underline{h}_1$ and $\underline{h}_2$ are the channel impulse response vectors from the transmitting antennas 1 and 2, respectively, and the superscript H on H^H and w^H indicates the Hermitian conjugation of H and w, respectively. If an impulse response length is assumed to be L, the channel impulse response vector is expressed as follows.

$$\underline{h}_i = [h_{i1}, h_{i2}, ..., h_{iL}] \tag{3}$$

[0007] Therefore, in the case of two transmitting antennas, equation (1) is calculated based on the following algebraic calculation.

$$H = \begin{bmatrix} h_{11} & h_{21} \\ h_{12} & h_{22} \\ \vdots & \vdots \end{bmatrix} \quad , \quad \underline{w} = [w_1, w_2]^T \quad , therefore \quad Hw = \begin{bmatrix} h_{11}w_1 + h_{21}w_2 \\ h_{12}w_1 + h_{21}w_2 \\ \vdots \end{bmatrix}$$

[0008] At the time of handover, weight vector $\underline{w}$ that maximizes the following equation is calculated instead of equation (1).

$$P = \underline{w}^H (H^H{}_1 H_1 + H^H{}_2 H_2 + ...)w \tag{4}$$

[0009] In equation (4), H_k is a channel impulse response signal from the k-th base station.

[0010] Then, the feedback information extraction unit 12 on the transmitting side extracts w_2 (in this case, w_1=1 is assumed) transmitted from a mobile station, from an incoming signal and an amplitude/phase control unit 13 multiplies a data signal to be transmitted from the transmitting antenna 2 by W_2. In this way, the degradation of both the amplitude and phase of signals received from the transmitting antennas 1 and 2 that are received on the receiving side are corrected in advance and are transmitted from the transmitting side.

[0011] In W-CDMA, two methods are stipulated: mode 1 for quantizing weight coefficient w_2 into one bit and mode 2 for quantizing w_2 into four bits. In mode 1, since control is exercised by transmitting one bit of feedback information for each slot, control speed is high. However, since quantization is rough, accurate control is impossible. In mode 2, since control is exercised by transmitting four bits of information, more accurate control is possible. However, in mode 2, since only one bit can be transmitted for each slot and feedback information of one word is transmitted for every four slots, control cannot track fading in the case of a high fading frequency, and amplitude/phase characteristics degrade. As described above, if the signal transfer rate of an uplink channel from a mobile station to a transmitting station, for transmitting feedback information is restricted, control accuracy and fading track speed have an inverse relationship.

[0012] The Release-99 specification of W-CDMA standard does not take into consideration a case where more than two transmitting antennas are used so as to avoid the degradation of uplink channel transmission efficiency due to feedback information transmission. However, if the increase of feedback information or the degradation of update speed is allowed, the number of antennas can also be increased to three or more. In particular, currently a case where four transmitting antennas are used is being extensively researched and developed.

[0013] If a closed-loop transmitting diversity system is applied to the radio base station of a cellular mobile communications system, a signal from each transmitting antenna independently suffers from fading, and ideally the same phase combination is performed at the antenna position of the mobile station. Therefore, a diversity gain corresponding to the number of transmitting antennas can be obtained and the gain can also be improved by the combination. Accordingly, the receiving characteristic is improved and the number of users accommodated in one cell can also be increased. "Ideally" means a case where there is neither transmission error of feedback information, control delay, channel response estimation error nor quantization error of a control amount. In reality, the characteristic degrades due to these factors compared with that of the ideal case.

[0014] In order to obtain a diversity gain corresponding to the number of antennas, antenna spacings (the distances between antennas) must be wide so that fading correlation may become sufficiently low. Generally, in order to suppress fading correlation to a sufficiently low level in the radio base station of a cellular mobile communications system, antenna spacings must be approximately 20 wavelengths. Since one wavelength is approximately 15cm in a 2GHz band, antennas must be installed approximately 3 meters apart. Therefore, if the number of transmitting antennas increases, an area needed to install antennas becomes wide and it becomes difficult to install antennas on the roof of a building and the like, which is a problem. Diversity gain is saturated as the number of transmitting antennas increases. Therefore, when the number of transmitting antennas reaches a specific value, the diversity gain cannot be improved any further even if the number of transmitting antennas is further increased.

[0015] When the number of transmitting antennas is increased, an amount of information to be fed back increases since feedback information must be transmitted to each antenna. Therefore, in that case, the transmission efficiency of an uplink channel degrades due to feedback information transmission or the control of transmitting diversity cannot track high-speed fading. As a result, the characteristic degrades, which is another problem.

Disclosure of Invention

[0016] An object of the present invention is to provide a transmitting diversity communications apparatus for suppressing the increase of uplink feedback information if the number of transmitting antennas is increased, suppressing the degradation of a characteristic in the case of a high fading frequency and requiring a small antenna installation space in the base-station.

[0017] The transmitting diversity communications apparatus of the present invention includes a transmitting diversity base station for controlling transmitting signals, according to information from a mobile station. The transmitting diversity communications apparatus comprises an antenna unit composed of a plurality of antenna groups, each consisting of a plurality of antennas, located close to each other so that the fading correlation between the antennas in the same group is high and groups are located apart from one another so that the fading correlation between the groups is low, and a control unit receiving both the first control information about intra-group antenna control with a low transfer rate that is transmitted from a mobile station and the second control information about inter-group antenna control and controlling the phase of a signal transmitted by the antenna unit.

[0018] According to the present invention, if signal control is applied to a closed-loop transmitting diversity system by the same method as in the conventional case where two transmitting antennas are used, by increasing the number of transmitting antennas, the tracking of fading fluctuations and transmitting-signal control performance can be prevented from degrading due to the increase of an amount of information to be transmitted from a mobile station to a base station.

[0019] In particular, according to the present invention, since the antenna unit of a base station is composed of a plurality of antenna groups each consisting of a plurality of antennas, and each intra-group antenna and each antenna group is set so that fading correlation is high within a group and so that fading correlation is low between groups,

respectively, only transmitting-signal control information between groups must be transmitted at a high speed from a mobile station to a base station and transmitting-signal control information within a group can be relatively slow. Therefore, transmitting diversity performance can be improved by effectively utilizing the limited transfer rate of an upward line from a mobile station to the base station.

Brief Description of Drawings

[0020]

Fig. 1 shows an example configuration of a transmitting diversity system using two transmitting antennas.
Fig. 2 shows the system configuration of the present invention.
Fig. 3 shows an example configuration of transmitting antennas of a base station according to the preferred embodiment of the present invention.
Fig. 4 shows the configuration of one preferred embodiment of the present invention.
Fig. 5 shows an example of a downlink pilot signal pattern in the preferred embodiment.
Fig. 6 shows both an example configuration of a base station transmitting antennas and antenna control information according to the preferred embodiment.
Fig. 7 shows an envelope correlation coefficient obtained when the angle dispersion $\Delta\phi$ of an input signal observed at a base station in a macro-cell environment is approximately 3.
Fig. 8 shows an example of the transmission format of feedback information in the preferred embodiment (No. 1) .
Fig. 9 shows an example of the transmission format of feedback information in the preferred embodiment (No. 2).
Fig. 10 shows an example of the transmission format of feedback information in the preferred embodiment (No. 3).
Fig. 11 shows an example of the transmission format of feedback information in the preferred embodiment (No. 4).
Fig. 12 shows an example configuration of a mobile station for transmitting feedback information to a base station according to the formats shown in Figs. 8 through 11.
Fig. 13 shows an example configuration of a base station in the second preferred embodiment of the present invention.
Fig. 14 shows an antenna phase difference control method within a group in the second preferred embodiment.
Fig. 15 shows the configuration of the third preferred embodiment of the present invention.

Best Mode for Carrying Out the Invention

[0021] The present invention relates to a closed-loop transmitting diversity method according to which the radio base station of a cellular mobile communications system is provided with a plurality of antennaslements, both different amplitude and phase control are exercised over the same transmitting data, according to feedback information from a mobile station and a plurality of pieces of data are transmitted using different antennas. On the mobile station side, the amplitude/phase control amounts are determined using a downward pilot signal; feedback information indicating the amplitude/phase control amounts are multiplexed with an uplink channel signal; and the data is transmitted to the base station.
[0022] Fig. 2 shows the system configuration of the present invention.
[0023] The pilot signal generation unit 20 of a base station generates N mutually orthogonal pilot signals $P_1(t)$, $P_2(t)$, ..., $P_N(t)$ and the pilot signals are transmitted using different antennas. N is the number of transmitting antennas. The following relationship is established between these pilot signals.

$$\int P_i(t)P_j(t)dt = 0 \quad (i \neq j)$$

[0024] Each pilot signal suffers from both amplitude and phase fluctuations due to fading, and a signal obtained by combining these pilot signals is transmitted to the receiving antenna 22 of a mobile station. The receiver of the mobile station estimates the channel impulse response vectors h_1, h_2, ..., h_N of each pilot signal by calculating the correlation between the incoming pilot signal and each of $P_1(t)$, $P_2(t)$, ..., P_N (t).
[0025] A control amount calculation unit 23 calculates and quantizes the amplitude/phase control vector (weight vector) $\underline{w}=[w_1, w_2, ..., w_N]^T$ of each transmitting antenna of the base station that maximizes power P expressed by equation (5)(the same as equation (1)) using these channel impulse response vectors. A multiplex unit 24 multiplexes the quantized vector with an upward channel signal as feedback information and transmits the signal to the base station side. However, in this case it is acceptable to transmit values w_2, w_3, ..., w_N obtained by assigning w_1=1.

$$P = \underline{w}^H H^H H \underline{w} \tag{5}$$

$$H = [\underline{h}_1, \underline{h}_2, .., \underline{h}_N] \tag{6}$$

[0026] In equation (6), $\underline{h}_1$ is a channel impulse response vector from transmitting antenna i. If an impulse response length is assumed to be L, $\underline{h}_1$ is expressed as follows.

$$\underline{h}_1 = [h_{i1}, h_{i2}, ..., h_{iL}]^T \tag{7}$$

[0027] At the time of hand-over, weight vector w that maximizes the following equation is calculated instead of equation (5).

$$P = \underline{w}^H (H_1^H H_1 + H_2^H H_2 + ...) \underline{w} \tag{8}$$

[0028] In equation (8), H_k is a channel impulse response signal from the k-th base station, and is the same as H_k in equation (4).

[0029] The multiplex unit 24 of the mobile station multiplexes the weight vector obtained in this way with an upward transmitting data signal and the vector is transmitted to the receiving antenna of the base station. In the base station, a feedback information extraction unit 25 extracts the feedback information received by a receiving antenna, and an amplitude/phase control unit 26 controls both the amplitude and phase transmitted from each transmitting antenna using a weight vector included in the feedback information. When the base station transmits a signal, both the amplitude and phase of which have been controlled from a transmitting antenna 21, the mobile station receives the signal as if the fluctuations due to just fading of both the amplitude and phase were compensated for. Therefore, optimal reception is possible. Since fading changes as time elapses, both the generation and transmission of feedback information must happen in real time. However, since both the transmission format and transfer rate of an uplink data signal from a mobile station to a base station is predetermined, it takes too much time to transmit a lot of information. Therefore, the control cannot track the fading fluctuations. In order to track the fading fluctuations, the transmission rate of feedback information must be high. However, since the transmission rate of an uplink control channel is limited, if a plurality of pieces of new information are sequentially transmitted in a short cycle in order to control transmitting diversity at a high speed, an amount of information included in one time transmission becomes small (quantization becomes rough), and highly accurate control becomes impossible.

[0030] In the preferred embodiment of the present invention, each coefficient value in a weight vector is calculated and fed back in a different cycle instead of calculating and feeding back a signal transmitted from each antenna in the same cycle.

[0031] The details are described below.

[0032] Fig. 3 shows an example configuration of the transmitting antennas of a base station according to the preferred embodiment of the present invention.

[0033] As shown in Fig. 3, in a base station, transmitting antennas compose a plurality of groups, each consisting of a plurality of antennas. Transmitting antennas in the same group are located close to one another so that the fading correlation between the antennas is high and groups are installed apart from one another so that the fading correlation between the groups is low. Fading correlation is a numeric value indicating how similarly two signals transmitted from different antennas fade when the signals are received on a receiving side. Doppler effect and the like cause fading by reflection on buildings and mobile objects. Therefore, if a plurality of antennas transmitting signals are located close to one another, it follows that a mobile station receives the respective signals through similar routes. Accordingly, the signals suffer from similar fading. In such a case, it is said that the fading correlation between the signals is high. If a plurality of antenna transmitting signals are located apart from one another, it follows that the respective signals take different routes to a mobile station receives the signals. Therefore, the signals fade differently and then are received by the mobile station. In such a case, it is said that the fading correlation between the signals is low.

[0034] In a mobile station, an antenna control amount between groups is calculated in a shorter cycle than that of the antenna control amount within a group, and is transmitted to a base station side as feedback information. Signals from base-station transmitting antennas in the same group have a high fading correlation; the signals suffer from almost the same fading, but the signals each have a phase difference depending on the angle at which the signals reach the receiving antenna of the mobile station. Therefore, each channel response estimation value estimated using the signals

from the plurality of base-station transmitting antennas in the same group has a phase difference that depends on the angle of the mobile station against the base station. Although these values change as the mobile station travels, the values change slowly compared with fading fluctuations. One antenna in each group is designated as a reference antenna, and each of the control amounts of antennas other than the reference antenna in the relevant group is normalized by the control amount of this reference antenna (each relative value calculated using the control amount of this reference antenna as a reference is used). This normalized antenna control amount in the group changes slowly as the mobile station travels. Therefore, the control cycle can be made relatively long.

[0035] However, since respective signals from base-station transmitting antennas belonging to different groups have a low fading correlation, the signals fade differently and independently by the time they reach the receiving antenna of the mobile station. Therefore, respective channel response estimation values (channel impulse response vector) estimated using respective signals from respective reference values belonging to different groups change quickly due to respective independent fading fluctuations. An antenna control amount obtained by normalizing the reference antenna control amount of one specific group by the reference antenna control amount of another group is defined as an inter-group antenna control amount. Since each inter-group antenna control amount changes quickly due to each independent fading fluctuation, in order to accurately control antennas, the control must be exercised in a short cycle.

[0036] The mobile station must recognize which signal comes from which group. However, it is sufficient to relate each antenna to each pilot signal transmitted from the antenna in advance. Since pilots are mutually orthogonal to one another, a receiving side can accurately recognize from which antenna the signal is transmitted by checking the pilot signal.

[0037] Both the inter-group antenna control amount $F_{1,m}$ and intra-group antenna control amount $G_{m,k}$ shown in Fig. 3 are calculated as follows. In the description given above, N, M and K=N/M are the total number of antennas, the number of antenna groups and the number of antennas in each group, respectively. * represents complex conjugation.

Overall reference antenna: Antenna #1

Intra-group reference antenna: Antenna # ((m-1) K+1) (m-1, ..., M)

$$F_{1,m} = \frac{w_{(m-1)K+1}}{w_1} \quad (m = 1,\ldots,M) \tag{9}$$

$$G_{m,k} = \frac{w_{(m-1)K+k+1}}{w_{(m-1)K+1}} \quad (m = 1,\ldots,M, k = 1,\ldots K) \tag{10}$$

[0038] Since fading correlation is high within a group, $|G_{m,k}| = 1$ can be assigned. Specifically, it can be considered that the change due to fading is small within a group, and it is sufficient to take into consideration only change in phase. In order to keep total transmission power constant (=1.0), $F_{1,m}$ must be normalized as follows.

$$F'_{1,m} = \frac{F_{1,m}}{\sqrt{\frac{1}{KM}\sum_{j=1}^{M}\left|F_{1,j}\right|^2}}$$

[0039] Next, the fluctuation rate of fading is described.

[0040] Fading fluctuation rate is expressed by Doppler frequency.

$$f_d = \frac{v}{\lambda}$$

[0041] In the equation described above, v is the travel speed of a mobile station and λ is the carrier wavelength. For example, if a carrier frequency is 2GHz and the travel speed of a mobile station is 60km/h, f_d becomes approximately 111Hz. However, the angle of arrival of an incoming wave changes as the mobile station travels. For example, if the mobile station travels at a speed of 200km/h at a place 200 meters ahead, the input angle changes by approximately 15 degrees per second. In this way, the fading fluctuation rate is higher by several tens of times to several hundreds of times than the fluctuation rate of an input angle. According to W-CDMA standards, a slot length is 666.7μs and the

update speed of feedback information is 1500Hz. Therefore, if information about fading is not updated for each slot, a track characteristic degrades. However, there is no need to feedback information about input angle for each slot. For example, there will be no problem if information is updated for every 15 slots (=one frame) .

[0042] By utilizing the difference in the fluctuation rates of the control information described above, a feedback amount of information can be reduced without performance degradation. Specifically, an inter-group antenna control amount changing at a high speed is updated and fed back in a short cycle, while each intra-group antenna control amount changing slowly compared with the inter-group antenna control amount is updated and fed back in a longer cycle. In other words, since the change of inter-group diversity control with a low fading correlation is faster than that of the data speed of feedback information, the frequency of updates is made large. However, since the change of intra-group diversity control with a high fading correlation is slower than that of the data speed of feedback information, the frequency of updates is made small.

[0043] Since each intra-group antenna control amount has been related to the angle of the mobile station with respect to the base station, in a macro-cell system with a relatively large cell radius, the deviation of an input angle becomes negligibly small. Therefore, a specific intra-group antenna control amount can also be used as the intra-group antenna control amount of another group. Specifically, transmitting only the intra-group control information of one specific group and controlling the antennas in the other group using this information can further reduce an amount of feedback information.

[0044] Fig. 4 shows the configuration of one preferred embodiment of the present invention.

[0045] A case where the number of antennas N=4 and the number of antenna groups M=2 is described. Apilot signal generation unit 30 generates N=4 pilot signals $P_1(t)$, $P_2(t)$, $P_3(t)$ and $P_4(t)$, and each of the signals is transmitted from one of transmitting antennas 31. These pilot signals use mutually orthogonal bit sequences.

[0046] Each transmitting antenna 31 transmits the pilot signal to a mobile station. In the mobile station, a receiving antenna 32 receives the four pilot signals transmitted from each of four transmitting antennas, and a control amount calculation unit 33 estimates the channel of signals transmitted from each transmitting antenna 31 using the respective pilot signal. As a result, the channel impulse response vector is obtained from each signal and a weight vector that maximizes equation (5) is calculated. Since a method for calculating this weight vector is already publicly known, the description is omitted. When the weight vector is calculated, the control amount calculation unit 33 transfers the vector to a multiplex unit 34 as feedback information. The multiplex unit 34 multiplexes the feedback information with an upward data signal and transmits the information from a transmitting antenna 35. In a base station, a receiving antenna 36 receives the signal from the mobile station, and a feedback information extraction unit 37 extracts the feedback information from the signal. The extracted feedback information is inputted to an amplitude/phase control unit 38, each weight coefficient W_1, W_2 and w_3 included in the feedback information is multiplied to the respective downward transmitting data signal of each corresponding antenna, and the transmitting antennas 31 transmit the downward transmitting data signals. In this way, in this preferred embodiment, a closed loop for performing transmitting diversity control, including a base station and a mobile, is implemented.

[0047] Fig. 5 shows examples of a downlink pilot signal pattern in this preferred embodiment.

[0048] If each corresponding code is multiplied by each of the pilot signals P_1 through P_4 shown in Fig. 5, and the products of the entire pilot signal pattern are added up the result "0" is obtained. Specifically, the pilot signals P_1 through P_4 form a mutually orthogonal code word.

[0049] Each pilot signal's amplitude and phase change independently due to fading, and the combination of these signals is received by the antenna of a mobile station. A mobile-station receiver can calculate the channel response estimation values h_1, h_2, h_3 and h_4 of each pilot signal by correlating the incoming pilot signals with corresponding pilot signals $P_1(t)$, $P_2(t)$, $P_3(t)$ and $P_4(t)$, respectively, that are stored in advance on the mobile station side and by averaging the obtained correlations.

[0050] Fig. 6 shows both an example configuration of base-station transmitting antennas according to this preferred embodiment and antenna control information thereof.

[0051] Fig. 6A shows the transmitting antenna configuration of a base station. It is assumed that antennas ANT1 and ANT2 form group 1, and antennas ANT3 and ANT4 form group 2. It is also assumed that antennas ANT1 and ANT3 are the reference antenna of groups 1 and 2, respectively. It is further assumed that antenna ANT1 is also the reference antenna of all the groups 1 and 2. Antennas ANT1 and ANT2 are located apart from each other by one wavelength. Antennas ANT3 and ANT4 are also located apart from each other by one wavelength. Antennas ANT1 and ANT3 are located apart from each other by 20 wavelengths. Antennas ANT2 and ANT4 are also located apart from each other by 20 wavelengths.

[0052] Here, the spatial correlative characteristic of a base-station antenna is described.

[0053] If the input angles of signals from mobile stations are uniformly distributed with dispersion $\Delta\phi$, the envelope correlation coefficient of input waves is expressed as follows. In the equation, d represents the distance between two antennas.

$$\rho = \left(\frac{\sin X}{X}\right)$$

$$X = \frac{\pi d \Delta\phi}{\lambda}$$

[0054] The angle dispersion $\Delta\phi$ of each input signal observed at the base station in a macro-cell environment is approximately 3 degrees. Fig. 7 shows the envelope correlation coefficient in this case. It is seen from Fig. 7 that at $d \cong 19\lambda$ the input signals become uncorrelated. Therefore, according to the present invention, fading correlation can be made low by setting the distance between antenna groups to approximately 19 wavelengths or more. Fading correlation can also be made high by setting the distance between antennas in each group to one wavelength or less.

[0055] However, fading correlation is affected by a variety of factors, such as the height at which the antenna is installed, the size of the antenna and the like. Therefore, it is acceptable if the antennas are installed so that the distance between any two antennas in the same group is approximately the wavelength of an incoming signal. However, a person having ordinary skill in the art should set the distance between groups so that fading correlation is almost "0" in any situation.

[0056] Description will return to Fig. 6. In the following description it is assumed that amplitude is not controlled and only phase is controlled. Specifically, only a phase amount ϕ_i is controlled by assigning $a_1=1$ to $w_i=a_i e^{j\phi_i}$. As shown in Fig. 6B, each of the control amount ϕ_1 of antenna ANT2 using antenna ANT1 as a reference, the control amount ϕ_2 of antenna ANT4 using antenna ANT3 as a reference and the control amount ϕ_3 of antenna ANT3 using antenna ANT1 as a reference is quantized and is transmitted to the base station as feedback information. If each of the control amounts is quantized using one bit, for example, the setting is as follows.

$$\begin{aligned} &\frac{-\pi}{2} < \phi_i \le \frac{\pi}{2} \Rightarrow \phi_i^Q = 0 \\ &\frac{\pi}{2} < \phi_i \le \frac{3\pi}{2} \Rightarrow \phi_i^Q = \pi \end{aligned} \qquad (11)$$

[0057] In the expression, ϕ_i^Q is a quantized control amount.

[0058] Figs. 8 through 11 show examples of the transmission format of feedback information in this preferred embodiment.

[0059] It is assumed that if $\phi_i^Q=0$, feedback information $b_i=0$ and that if $\phi_i^Q=\pi$, feedback information $b_i=1$. As shown in Fig. 8, this feedback information is multiplexed with an upward channel so that the transmission rate of b_3 may become higher than the transmission rate of b_1 or b_2 and is transmitted to a base station. One frame of length 10ms is composed of 15 slots in compliance with the W-CDMA frame format. This transmission format transmits feedback information of one bit in each slot. Format1 transmits both one b_1 and one b_2 in one frame, and format2 transmits both two b_1 and two b_2 in one frame.

[0060] In the base station, the phase control of each transmitting antenna is conducted using the feedback information received in an uplink channel. A corresponding antenna is directly controlled by the feedback information received in the immediately previous slot. In this case, antennas other than the corresponding antenna store the latest feedback information and use the information for their control.

[0061] However, ANT4 shown in Fig. 6A is controlled not only by control amount d2, but also by the control amount d3 of ANT 3. Specifically, ANT4 is frequently controlled by d3 and is also controlled by d2 less frequently. This description also applies to ANT4 shown in Fig. 6B.

[0062] Filtering feedback information can also reduce the number of transmission errors and the number of quantization errors. For example, for the filtering, a method using the average value of the control amount of the feedback information received in the immediately previous slot and the control amount of the feedback information received in receiving slots before the immediately previous slot is used.

[0063] As the feedback information of an intra-group antenna control amount, an updated control amount is transmitted every time the feedback information is transmitted. However, for example, alternatively, in the same frame, the same feedback information can also be repeatedly transmitted. In this case, the number of transmission errors in the base station can be reduced by combining a plurality of pieces of feedback information received in the frame.

[0064] Since each intra-group antenna control amount relates to the angle of a mobile station against the base station

in a macro-cell system with a to some extent large cell radius, the deviation of the input angle within a group is negligibly small. Therefore, if control is exercised within each group using the same intra-group antenna control amount, there is no problem. Therefore, transmitting only the intra-group control information of one specific group and controlling other groups using this information can further reduce an amount of feedback information.

[0065] Fig. 9 shows a feedback information transmission format used to transmit only b_1 as intra-group control information. Format3 transmits two b_1 in one frame and format 4 transmits four b_1 in one frame.

[0066] In this preferred embodiment too, as the feedback information of an intra-group antenna control amount, an updated control amount can be transmitted every time the feedback information is transmitted. Alternatively, for example, the same feedback information can be repeatedly transmitted within the same frame.

[0067] Another transmission format in which a control amount calculated in a mobile station is quantized using a plurality of bits is described below.

[0068] Fig. 10 shows the feedback information transmission format in which b_1 and b_3 are quantized using three bits and four bits, respectively. Tables 1 and 2 of Fig. 11 show the correspondence between the feedback information b_3 of an inter-group antenna control amount and a control amount. Table 3 shows the correspondence between the feedback information b_1 of an intra-group antenna control amount and a control amount.

[0069] In this example, only the feedback information of an intra-group antenna control amount b_1 is transmitted using the format shown in Fig. 9. As is clearly seen from Tables 1 and 2 of Fig. 11, feedback information bit b_3 is composed of four bits; three bits of $b_3(3)$ through $b_3(1)$ representing a phase control amount and one bit of $b_3(0)$ representing an amplitude control amount. Format5 shown in Fig. 10 includes feedback information bit b_3 in one frame. However, three words of feedback information bit b1 are composed of three bits of b_1 (2) through b_1 (0) representing a phase control amount. According to formats shown in Fig. 10, three bits of feedback information bit b_1 are distributed and located in one frame, and all the three bits together form one word.

[0070] Fig. 12 shows an example configuration of a mobile station that transmits feedback information to a base station according to the formats shown in Figs. 8 through 11.

[0071] On receipt of a signal from a base station via its receiving antenna, a mobile station branches the receiving signal into two signals and inputs one signal and the other signal to a data channel despreading unit 41 and a pilot channel despreading unit 44, respectively. The data channel dispreading unit 41 despreads the data channel signal and inputs the signal to both a channel estimation unit 42 and a receiver 43. The receiver 43 reproduces the downlink data signal, based on the channel estimation result of the channel estimation unit 42 and presents the signal to a user as voice or data. The pilot channel dispreading unit 44 despreads the incoming signal using a pilot channel dispreading code and inputs the signal to a channel estimation unit 45. The channel estimation unit 45 correlates the despread signal to each pilot signal pattern and obtains channel estimation values $H=[h_1, h_2, h_3$ and $h_4]$ for paths from each transmitting antenna to the mobile station. A control amount calculation unit 46 calculates a weight vector based on these channel estimation values and determines feedback information to be transmitted. A multiplex unit 47 multiplexes this feedback information with an upward control channel. A data modulation unit 48 modulates the feedback information. A spreading modulation unit 49 spread-modulates the feedback information. Then, the feedback information is transmitted to the base station from a transmitting antenna 50.

[0072] In Fig. 13, the same reference numbers are attached to the same constituent components as those in Fig. 4 and their descriptions are omitted.

[0073] In this preferred embodiment, a base station uses both uplink feedback information and an uplink channel arriving method estimation result as intra-group antenna control information. In the base station, input direction estimation units 62 and 63 estimate the arriving direction of an incoming signal based on an uplink receiving signal received by an array antenna (a plurality of antennaslements used in transmission diversity: transmitting/receiving antenna 60). Since arriving direction strongly depends on the angle of a mobile station against a base station, a method for setting the direction of a downlink transmitting beam (direction in which the strength of a wave transmitted from an antenna is large) to this uplink signal input direction is known. However, in a system where uplink and downlink frequencies are different, this assumption does not always hold true and depends on the propagation environment.

[0074] Upon receipt of the uplink feedback information via an antenna 60, a receiving processing unit 61 performs the despeading and the like of the uplink feedback information and relays the information to a feedback information extraction unit 37. When the feedback information extraction unit 37 extracts a control amount from the uplink feedback information, an amplitude/phase control unit 38' compares the control amount with the arriving direction estimation value and determines to use either the control amount received from the uplink line or the arriving direction estimation value. Then, the unit 38' controls the amplitude/phase of a transmitting signal.

[0075] As shown in Fig. 14, in this preferred embodiment, if the intra-group phase difference is not within a specific range $[\theta-\Delta,\theta-\Delta]$ with the arriving direction estimation result θ of the uplink channel as a center, control is exercised using only the arriving direction estimation result θ since the control amount by the upward feedback information is related to an uplink channel arriving direction estimation result θ. Specifically, if the control amount in the feedback information is too far from the arriving direction estimation result θ, it is judged that a bit error or the like has occurred

during transmission of the feedback information, and the feedback information is inaccurate. Then, the feedback information is discarded and only phase is controlled using the arriving direction estimation result θ.

[0076] Alternatively, a control amount in the uplink feedback information of intra-group phase difference information can be sampled for a prescribed time period. If it is judged that variance of the samples is large (for example, specifically, if the samples are dispersed more widely than a specific predetermined threshold value), control can be exercised using only the arriving method estimation result θ without utilizing the feedback information.

[0077] Fig. 15 shows the configuration of the third preferred embodiment of the present invention.

[0078] In Fig. 15, the same reference numbers are attached to the same components as those in Fig. 4, and their descriptions are omitted.

[0079] In this case, the transmitting powers of pilot signals P_1 and P_3 are set smaller than the transmitting powers of pilot signals P_2 and P_4, respectively. In this preferred embodiment, this is implemented by multiplying pilot signals P_2 and P_4 by a coefficient α ($0<\alpha<1$). Although pilot signals P_2 and P_4 are needed to estimate channel impulse response vectors $\underline{h}_2$ and $\underline{h}_4$, $\underline{h}_2$ and $\underline{h}_4$ have high fading correlations to $\underline{h}_1$ and $\underline{h}_3$, respectively. Therefore, $\underline{h}_2/\underline{h}_1$ and $\underline{h}_4/\underline{h}_3$ that are normalized by them strongly depend on an angle of the mobile station against the base station. Since these values fluctuate slowly compared with fading fluctuation, estimation accuracy can be improved by taking a long time average of pilot signals P_2 and P_4 even if incoming power on the mobile station side is low. Both ϕ_1 and ϕ_2 are calculated as follows.

$$\phi_1 = \underline{h}_2/\underline{h}_1,\ \phi_2 = \underline{h}_4/\underline{h}_3 \tag{12}$$

[0080] Since interference to data signals by pilot signals can be suppressed to a low level by setting the transmitting powers of pilot signals P_2 and P_4 to a low level, transmission capacity can be increased.

[0081] Since both $\underline{h}_2/\underline{h}_1$ and $\underline{h}_4/\underline{h}_3$ depend on the angle of the mobile station against the base station and fluctuate more slowly than a fading fluctuation, estimation accuracy can be improved by taking a long time average of pilot signals P_2 and P_4 even if an incoming power is low. For example, estimation values $\phi_1(n)$, $\phi_2(n)$ and ϕ_3 (n) at the n-th slot can be calculated as follows. In these equations, N is the estimated average number of slots of estimation values ϕ_1 (n) and $\phi_2(n)$.

$$\phi_1(n) = \frac{1}{N}\sum_{i=0}^{N-1}\frac{\underline{\underline{h}}_2(n-i)}{\underline{h}_1(n-i)}$$

$$\phi_2(n) = \frac{1}{N}\sum_{i=0}^{N-1}\frac{\underline{\underline{h}}_4(n-i)}{\underline{h}_3(n-i)}$$

$$\phi_3(n) = \frac{\underline{\underline{h}}_3(n)}{\underline{h}_1(n)}$$

[0082] In this way, when both ϕ_1 and ϕ_2 are calculated by taking a N-times time (number of slots) average of ϕ_3, the same estimation accuracy as that of ϕ_3 can be obtained even if α = 1/N. Specifically, in case N=4, α = 1/4 can be assigned.

Industrial Applicability

[0083] If the number of transmitting antennas is increased by utilizing differences in the fluctuation rate of control information, the following effects can be obtained.

- The increase in the amount of upward feedback information can be suppressed.
- Characteristics degrade little in the case of a high fading frequency.
- The antenna installation space of a base station can be reduced.

Claims

1. A transmitting diversity communications apparatus, including a base station adopting a transmitting diversity method, for controlling transmitting signals according to information from a mobile station, comprising:

   antenna means composed of a plurality of antenna groups, each group consisting of a plurality of antennas located close to one another so that fading correlation between the antennas is high, and the antenna groups are located apart from one another so that fading correlation between the groups is low; and
   control means for receiving both first control information for intra-group antenna control, with a low transfer rate and second control information for inter-antenna group control, with a high transfer rate that are transmitted from a mobile station, and controlling a phase of a signal transmitted by the antenna means.

2. The transmitting diversity communications apparatus according to claim 1, wherein the mobile station determines a control amount of the phase using pilot signals transmitted from the base station.

3. The transmitting diversity communications apparatus according to claim 1, wherein said control means also controls amplitude in addition to the phase.

4. The transmitting diversity communications apparatus according to claim 3, wherein the mobile station determines control amounts of both the phase and amplitude using pilot signals transmitted from the base station.

5. The transmitting diversity communications apparatus according to claim 4, wherein the mobile station estimates a channel response from each antenna to the mobile station by correlating a pilot signal from the base station to a known pilot signal on a mobile station side and calculating the control amount using this channel response estimation value.

6. The transmitting diversity communications apparatus according to claim 1, wherein the mobile station transmits information describing the difference in channel response estimation values between each intra-group antenna of said antenna means and a reference antenna and information describing the difference in channel response estimation values between each antenna group and a reference antenna of a specific antenna group to the base station as the first and second control information, respectively.

7. The transmitting diversity communications apparatus according to claim 1, wherein the mobile station transmits control information about each antenna group and control information about an intra-group antenna within a specific antenna group to the base station as the second and first control information, respectively.

8. The transmitting diversity communications apparatus according to claim 1, wherein said control means controls the transmitting of a signal from the base station using an input direction estimation result of an uplink channel signal in addition to the first and second control information.

9. The transmitting diversity communications apparatus according to claim 8, wherein if transmitting signal control amounts obtained from the first and second control information do not fall within a specific range with an arriving direction estimation result of the uplink channel signal as a center, transmission is controlled using the input direction estimation result.

10. The transmitting diversity communications apparatus according to claim 8, wherein if transmitting signal control amount dispersion obtained from the first control information is larger than a prescribed value, transmission is controlled using only an arriving direction estimation result.

11. The transmitting diversity communications apparatus according to claim 1, wherein control is exercised by a filtering result using both currently received first and second information and one or more previously received first and second control information.

12. The transmitting diversity communications apparatus according to claim 1, wherein the power of a signal transmitted from an antenna other than a reference antenna is set at a lower level than the power of a signal transmitted from a reference antenna of each antenna group.

13. A transmitting diversity communications apparatus, including a base station adopting a transmitting diversity meth-

od, for controlling transmitting signals according to information from a mobile station, comprising:

antenna means composed of a plurality of antenna groups, each group consisting of a plurality of antennas located close to one another so that fading correlation between the antennas is high, and the antenna groups are located apart from one another so that fading correlation between the groups is low; and
control means for receiving both first control information for intra-group antenna control, with a low transfer rate and second control information for inter-antenna group control, with a high transfer rate that are transmitted from a mobile station, and controlling both amplitude and phase of a signal transmitted by the antenna means.

14. A mobile station of a transmitting diversity communications apparatus, including a base station adopting a transmitting diversity method, for controlling transmitting signals according to information from a mobile station, comprising:

receiving means for receiving a signal transmitted from an antenna means composed of a plurality of antenna groups, each group consisting of a plurality of antennas located close to one another so that fading correlation between the antennas is high, and the antenna groups are located apart from one another so that fading correlation between the groups is low;
antenna specifying means for identifying an antenna that has transmitted the received signal; and
transmitting means for transmitting first control information about intra-group antenna control of the received signal to the base station at a prescribed transfer rate and transmitting second control information about inter-group antenna control of the received signal to the base station at a higher transfer rate than the prescribed transfer rate.

15. A transmitting diversity communications method, including a base station adopting a transmitting diversity method, for controlling transmitting signals according to information from a mobile station, comprising:

providing a plurality of antenna groups, each group consisting of a plurality of antennas, placing the antennas in the same group close to one another so that fading correlation between the antennas in the same group is high and placing the antenna groups apart from one another so that fading correlation between the groups is low; and
receiving both first control information for intra-group antenna control, with a low transfer rate and second control information for inter-antenna group control, with a high transfer rate that are transmitted from a mobile station and controlling the phase of a signal transmitted by the antenna unit.

16. A communications system for controlling the phase of each of transmitting signals transmitted from a plurality of antennas of a base station according to phase control information from a mobile station, wherein
some of the plurality of antennas of the base station have a location relation in which, with respect to one antenna, other antennas are placed where fading correlation is high, and where fading correlation is low, and
the mobile station transmits phase control information about an antenna located in the position having a high fading correlation and phase control information about an antenna located in the position having a low fading correlation to the base station with low frequency and high frequency, respectively.

17. A communications system for controlling the phase of each of transmitting signals transmitted from a plurality of antennas of a base station according to phase control information from a mobile station, wherein
some of the plurality of antennas of the base station have a location relation in which, with respect to one antenna, other antennas are placed where fading correlation is high, and where fading correlation is low, and
the mobile station transmits phase control information about an antenna located in the position having a high fading correlation to the base station with lower frequency than frequency of phase control information of an antenna located in the position having a low fading correlation.

18. A communications system for controlling the phase of each of transmitting signals transmitted from a plurality of antennas of a base station according to phase control information from a mobile station, wherein
all the plurality of antennas except a specific antenna of the base station are located in positions having a specific fading correlation to the antenna, and
the mobile station transmits phase control information about the antennas except the specific antenna to the base station with frequency corresponding to the specific fading correlation.

19. A communications system for controlling the phase of each of transmitting signals transmitted from a plurality of

antennas of a base station according to phase control information from a mobile station, wherein
all the plurality of antennas except a specific antenna of the base station are located in positions having a high fading correlation to the antenna, and
the mobile station transmits phase control information about the antennas except the specific antenna to the base station with low frequency.

20. A communications system for controlling the phase of each of transmitting signals transmitted from a plurality of antennas of a base station according to phase control information from a mobile station, wherein
all the plurality of antennas except a specific antenna of the base station are located in positions having a low fading correlation to the antenna, and
the mobile station transmits phase control information about all the antennas except the specific antenna to the base station with high frequency.

21. A mobile station of a communications system for controlling the phase of each of signals transmitted from a plurality of antennas on a base station side where some of the plurality of antennas and the other antennas except a specific antenna are located in positions having a high fading correlation and in positions having a low fading correlation, respectively, to the antenna according to phase control information from a mobile station, comprising:

control means for generating phase control information about the plurality of antennas; and
transmitting means for transmitting phase control information about an antenna located in a position having a high fading correlation and phase control information of an antenna located in a position having a low fading correlation to the base station with low frequency and high frequency, respectively.

22. A mobile station of a communications system for controlling the phase of each of signals transmitted from a plurality of antennas on a base station side where some of the plurality of antennas and the other antennas except a specific antenna are located in positions having a high fading correlation and in positions having a low fading correlation, respectively, to the antenna according to phase control information from a mobile station, comprising:

control means for generating phase control information about the plurality of antennas; and
transmitting means for transmitting phase control information about an antenna located in a position having a high fading correlation to the base station with lower frequency than frequency of phase control information of an antenna located in a position having a low fading correlation.

23. A mobile station of a communications system for controlling the phase of each of transmitting signals transmitted from a plurality of antennas on a base station side where all the plurality of antennas except a specific antenna are located in positions having a specific fading correlation to the antenna according to phase control information from a mobile station, comprising:

control means for generating phase control information about the plurality of antennas; and
transmitting means for transmitting the corresponding antenna phase control information to the base station with frequency corresponding to the fading correlation.

24. A mobile station of a communications system for controlling the phase of each of signals transmitted from a plurality of antennas on a base station side where the other of the plurality of antennas are located in positions having a high fading correlation with one antenna according to phase control information from a mobile station, comprising:

control means for generating phase control information about the plurality of antennas; and
transmitting means for transmitting antenna phase control information about all the antennas except the specific antenna to the base station with low frequency.

25. A mobile station of a communications system for controlling the phase of each of signals transmitted from a plurality of antennas on a base station side where all the plurality of antennas except a specific antenna are located in positions having a low fading correlation to the antenna according to phase control information from a mobile station, comprising:

control means for generating phase control information about the plurality of antennas; and
transmitting means for transmitting phase control information about all the antennas except a specific antenna to the base station with high frequency.

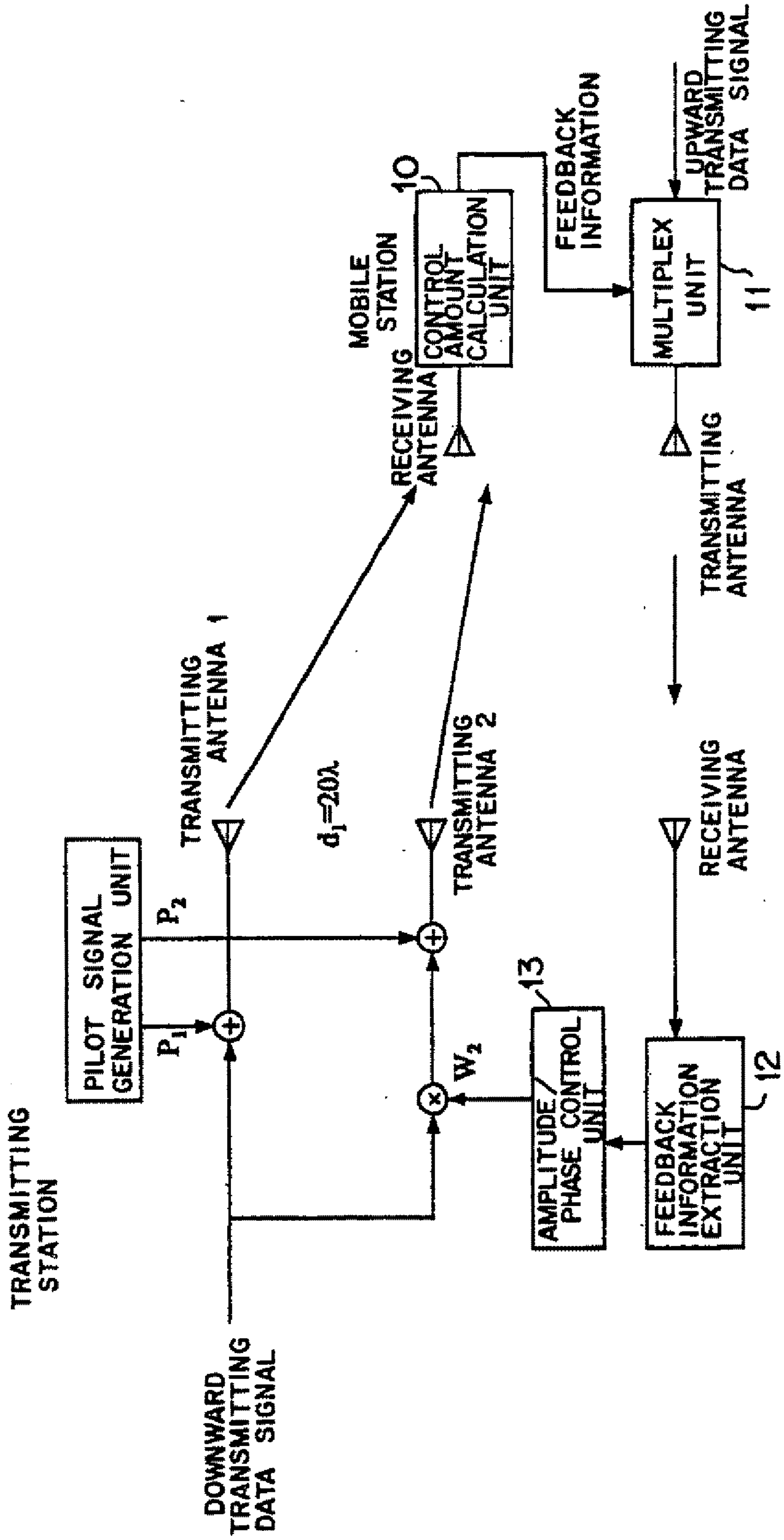
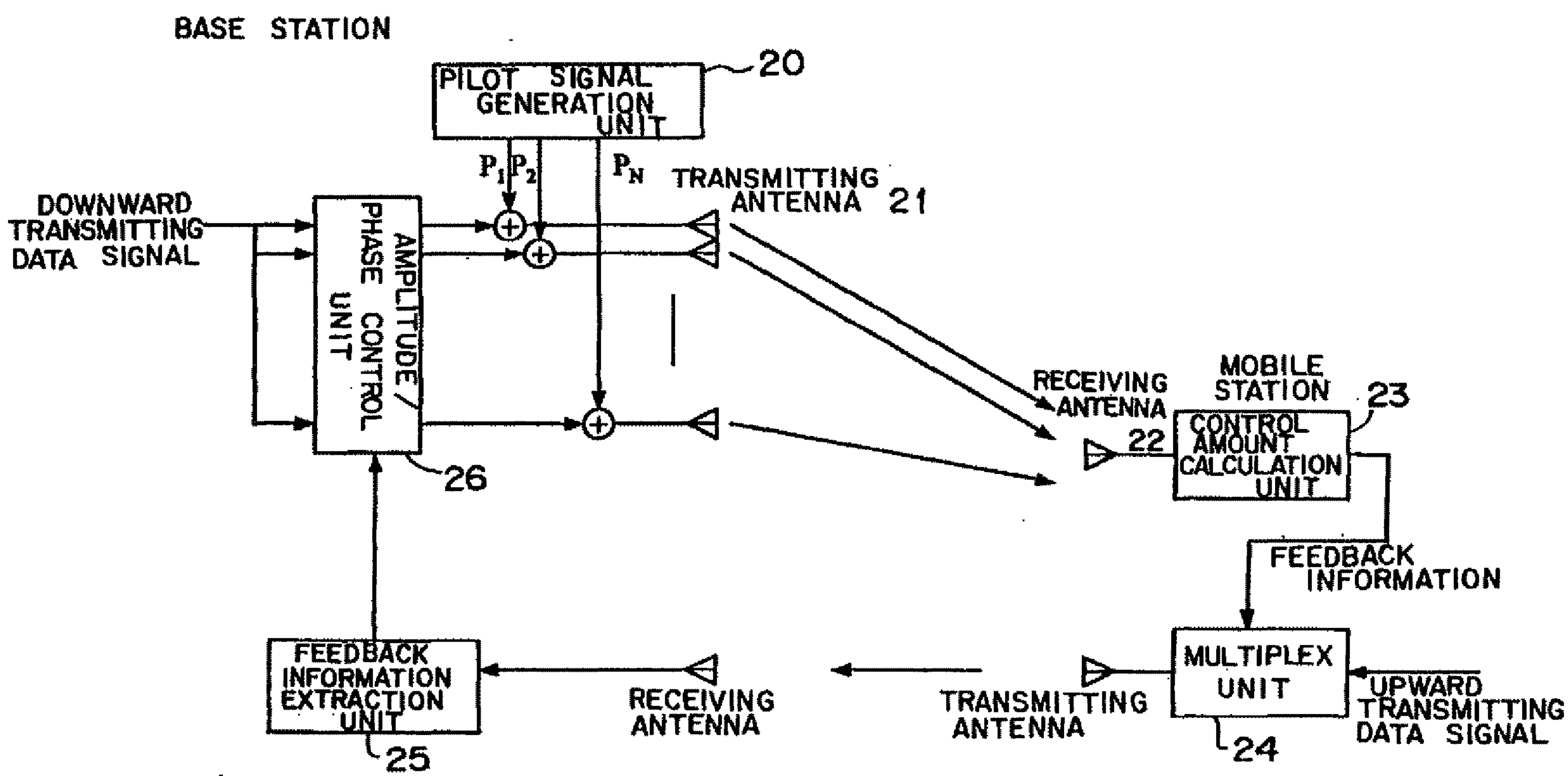

TRANSMITTING STATION
DOWNWARD TRANSMITTING DATA SIGNAL
PILOT SIGNAL GENERATION UNIT
P1
P2
TRANSMITTING ANTENNA 1
TRANSMITTING ANTENNA 2
d1=20λ
W2
AMPLITUDE/PHASE CONTROL UNIT
13
FEEDBACK INFORMATION EXTRACTION UNIT
12
RECEIVING ANTENNA
MOBILE STATION
RECEIVING ANTENNA
CONTROL AMOUNT CALCULATION UNIT
10
FEEDBACK INFORMATION
MULTIPLEX UNIT
11
UPWARD TRANSMITTING DATA SIGNAL
TRANSMITTING ANTENNA

FIG. 1


BASE STATION
PILOT SIGNAL GENERATION UNIT
20
P_1
P_2
P_N
TRANSMITTING ANTENNA
21
DOWNWARD TRANSMITTING DATA SIGNAL
AMPLITUDE / PHASE CONTROL UNIT
26
RECEIVING ANTENNA
22
MOBILE STATION
CONTROL AMOUNT CALCULATION UNIT
23
FEEDBACK INFORMATION
MULTIPLEX UNIT
24
UPWARD TRANSMITTING DATA SIGNAL
TRANSMITTING ANTENNA
RECEIVING ANTENNA
FEEDBACK INFORMATION EXTRACTION UNIT
25

F I G. 2

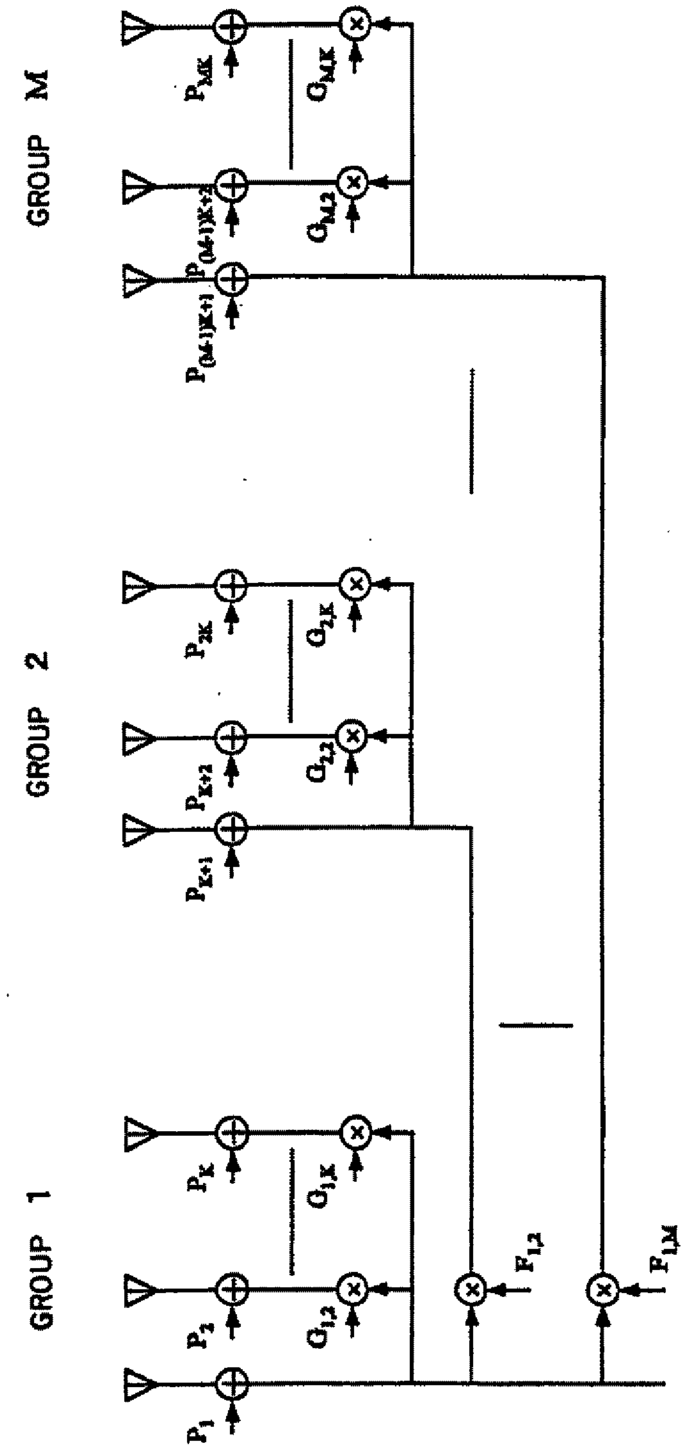
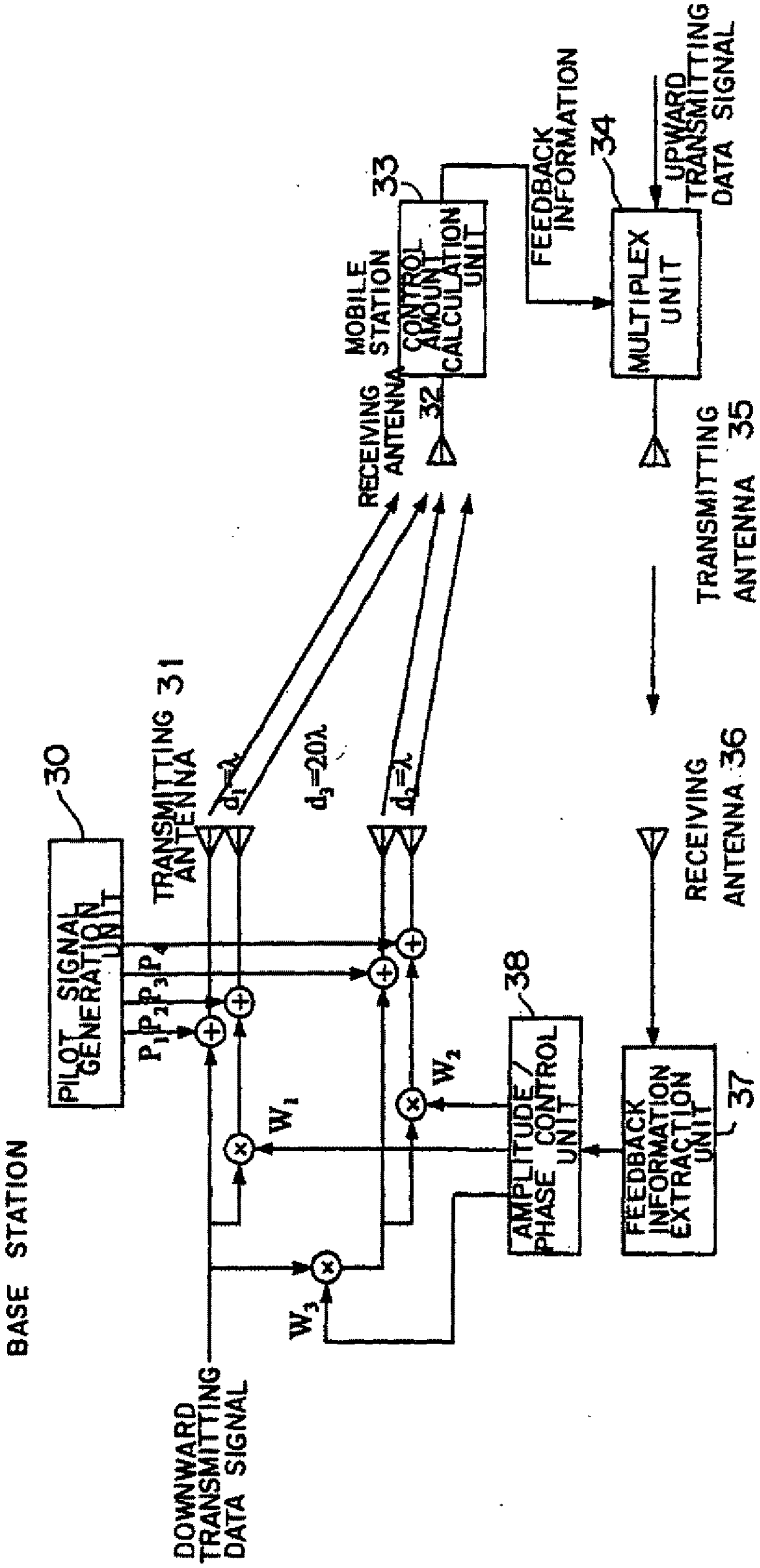

GROUP 1
GROUP 2
GROUP M
P_1
P_2
P_K
$G_{1,2}$
$G_{1,K}$
$F_{1,2}$
$F_{1,M}$
P_{K+1}
P_{K+2}
P_{2K}
$G_{2,2}$
$G_{2,K}$
$P_{(M-1)K+1}$
$P_{(M-1)K+2}$
P_{MK}
$G_{M,2}$
$G_{M,K}$

FIG. 3

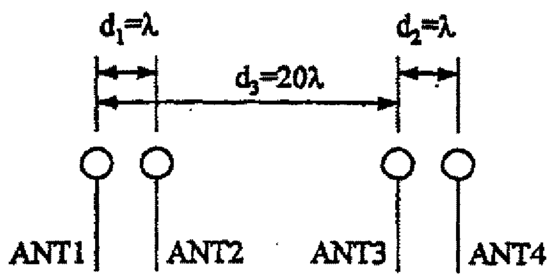

BASE STATION
DOWNWARD TRANSMITTING DATA SIGNAL
PILOT SIGNAL GENERATION UNIT
30
P_1 P_2 P_3 P_4
W_1
W_2
W_3
TRANSMITTING ANTENNA 31
$d_1=\lambda$
$d_3=20\lambda$
$d_2=\lambda$
AMPLITUDE / PHASE CONTROL UNIT
38
FEEDBACK INFORMATION EXTRACTION UNIT
37
RECEIVING ANTENNA 36
MOBILE STATION
RECEIVING ANTENNA
32
CONTROL AMOUNT CALCULATION UNIT
33
FEEDBACK INFORMATION
MULTIPLEX UNIT
34
UPWARD TRANSMITTING DATA SIGNAL
TRANSMITTING ANTENNA 35

F I G. 4

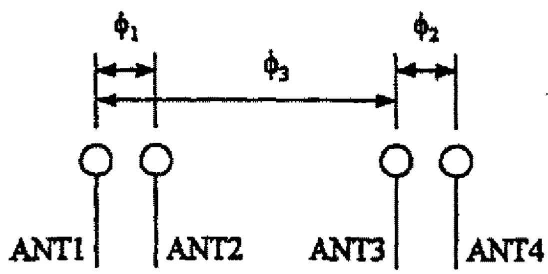
ANTENNA 1 P_1	A	A	A	A	A	A	A	A	A	A	A	A
ANTENNA 2 P_2	A	A	-A	-A	A	A	-A	-A	A	A	-A	-A
ANTENNA 3 P_3	A	-A	A	-A	A	-A	A	-A	A	-A	A	-A
ANTENNA 4 P_4	A	-A	-A	A	A	-A	-A	A	A	-A	-A	A

$A=1+j$

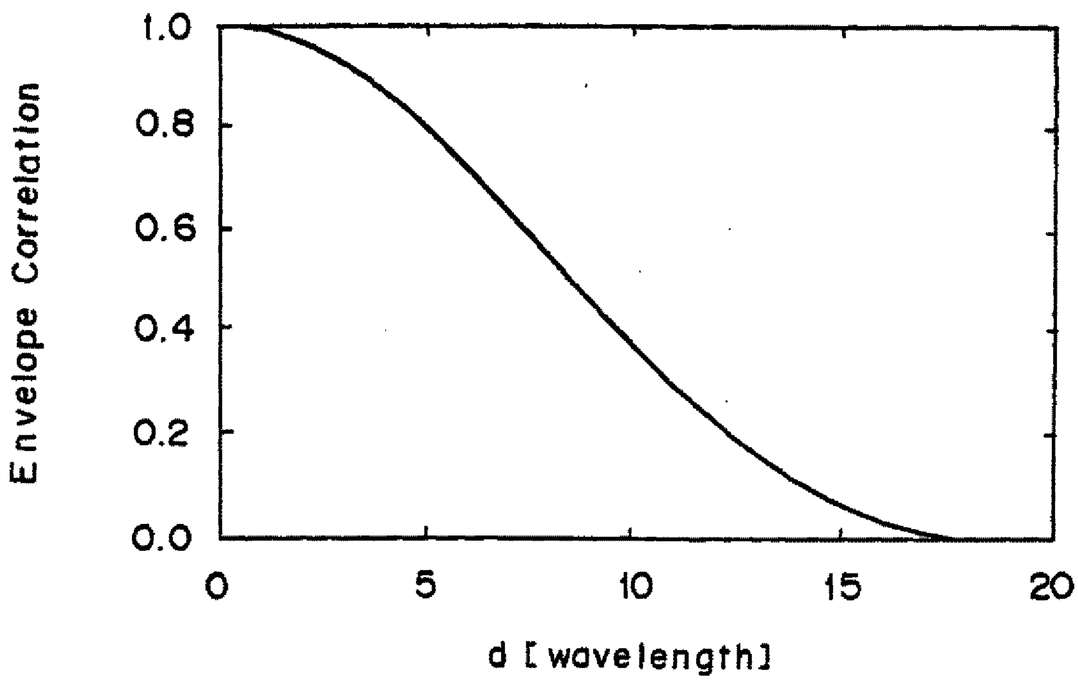
FIG. 5


$d_1=\lambda$
$d_2=\lambda$
$d_3=20\lambda$
ANT1
ANT2
ANT3
ANT4

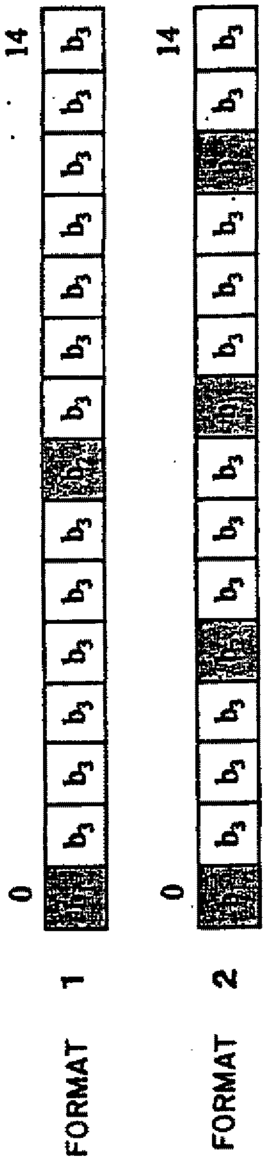
F I G. 6 A


ϕ_1
ϕ_2
ϕ_3
ANT1
ANT2
ANT3
ANT4

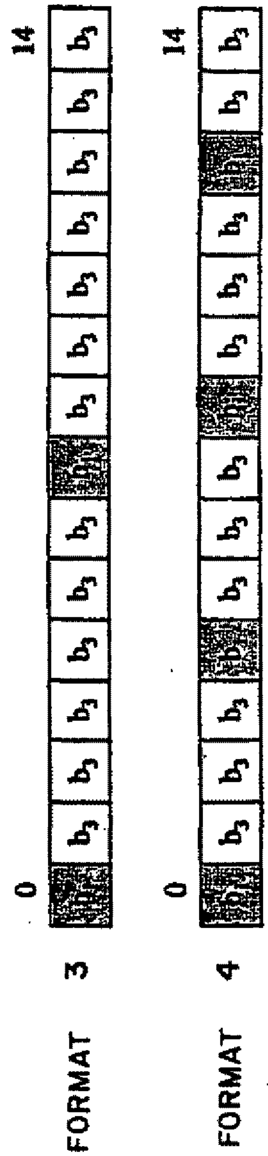
F I G. 6 B


1.0
0.8
0.6
0.4
0.2
0.0
Envelope Correlation
0
5
10
15
20
d [wavelength]

F I G. 7



F I G. 8



F I G. 9

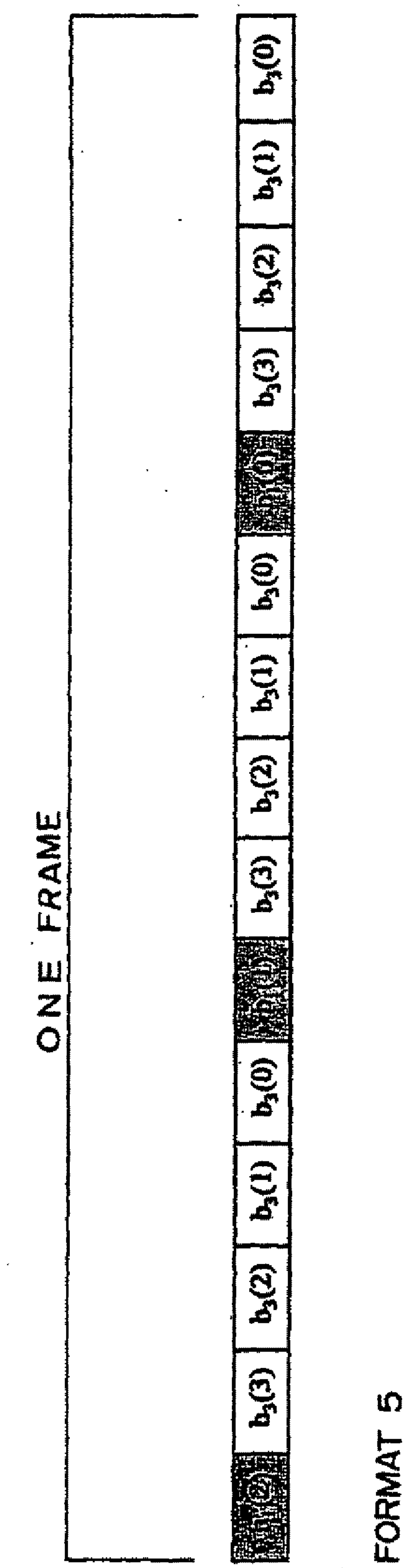

ONE FRAME
$b_3(3)$
$b_3(2)$
$b_3(1)$
$b_3(0)$
$b_3(3)$
$b_3(2)$
$b_3(1)$
$b_3(0)$
$b_3(3)$
$b_3(2)$
$b_3(1)$
$b_3(0)$
FORMAT 5

FIG. 10

TABLE 1. FEEDBACK BIT ($b_3(0)$)

$b_3(0)$	ANTENNA 1 AMPLITUDE	ANTENNA 2 AMPLITUDE
0	0.2	0.8
1	0.8	0.2

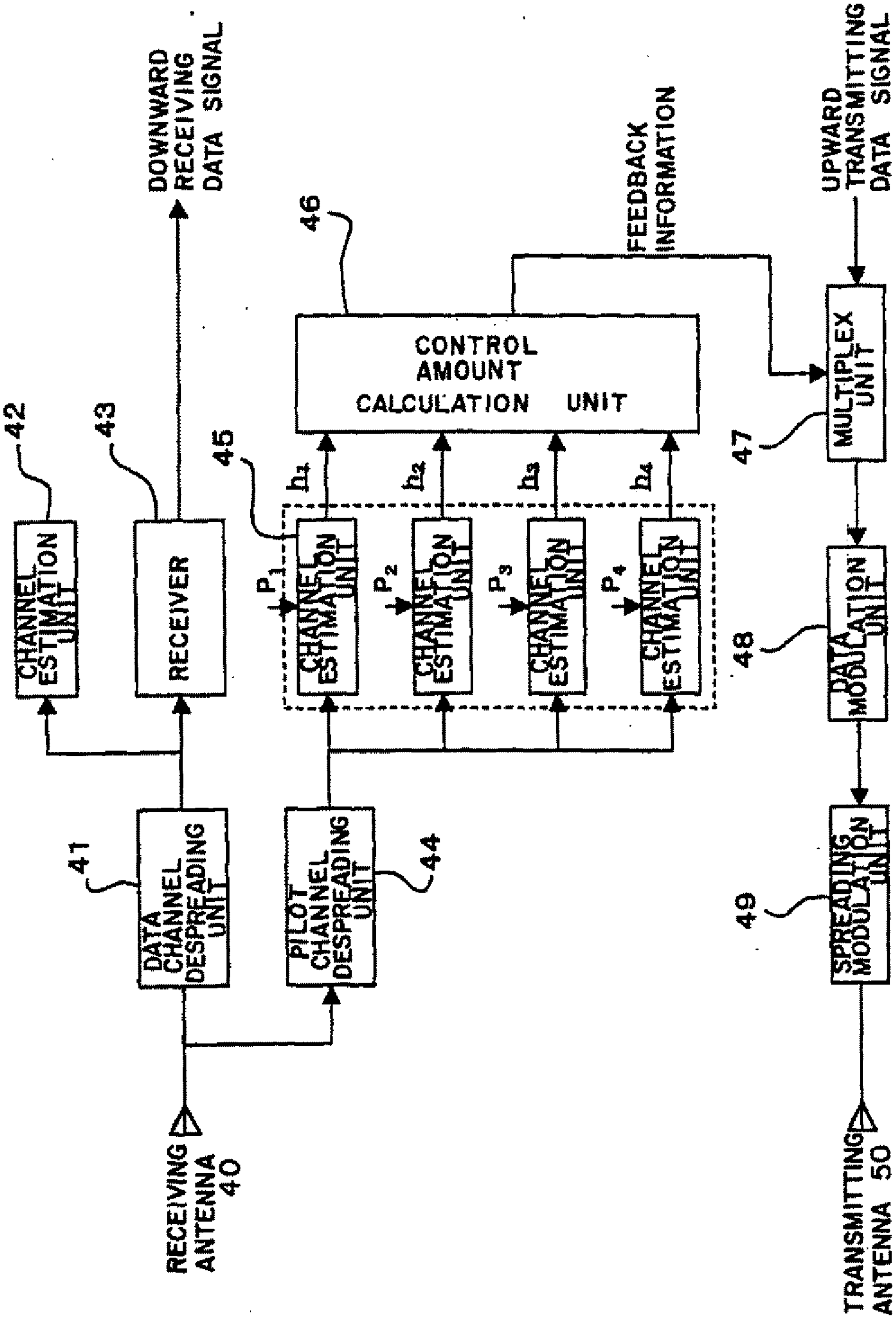
TABLE 2. FEEDBACK BIT ($b_3(3)$, $b_3(2)$, $b_3(1)$)

$b_3(3)$, $b_3(2)$, $b_3(1)$	INTER-ANTENNA PHASE DIFFERENCE (DEGREE)
000	180
001	-135
010	-90
011	-45
100	0
101	45
110	90
111	135

TABLE 3. FEEDBACK BIT ($b_1(2)$, $b_1(1)$, $b_1(0)$)

$b_1(2)$, $b_1(1)$, $b_1(0)$	INTER-ANTENNA PHASE DIFFERENCE (DEGREE)
000	180
001	-135
010	-90
011	-45
100	0
101	45
110	90
111	135

FIG. 11


RECEIVING ANTENNA 40
DATA CHANNEL DESPREADING UNIT
41
PILOT CHANNEL DESPREADING UNIT
44
CHANNEL ESTIMATION UNIT
42
RECEIVER
43
DOWNWARD RECEIVING DATA SIGNAL
45
P1
P2
P3
P4
CHANNEL ESTIMATION UNIT
h1
h2
h3
h4
CONTROL AMOUNT CALCULATION UNIT
46
FEEDBACK INFORMATION
UPWARD TRANSMITTING DATA SIGNAL
MULTIPLEX UNIT
47
DATA MODULATION UNIT
48
SPREADING MODULATION UNIT
49
TRANSMITTING ANTENNA 50

F I G. 12

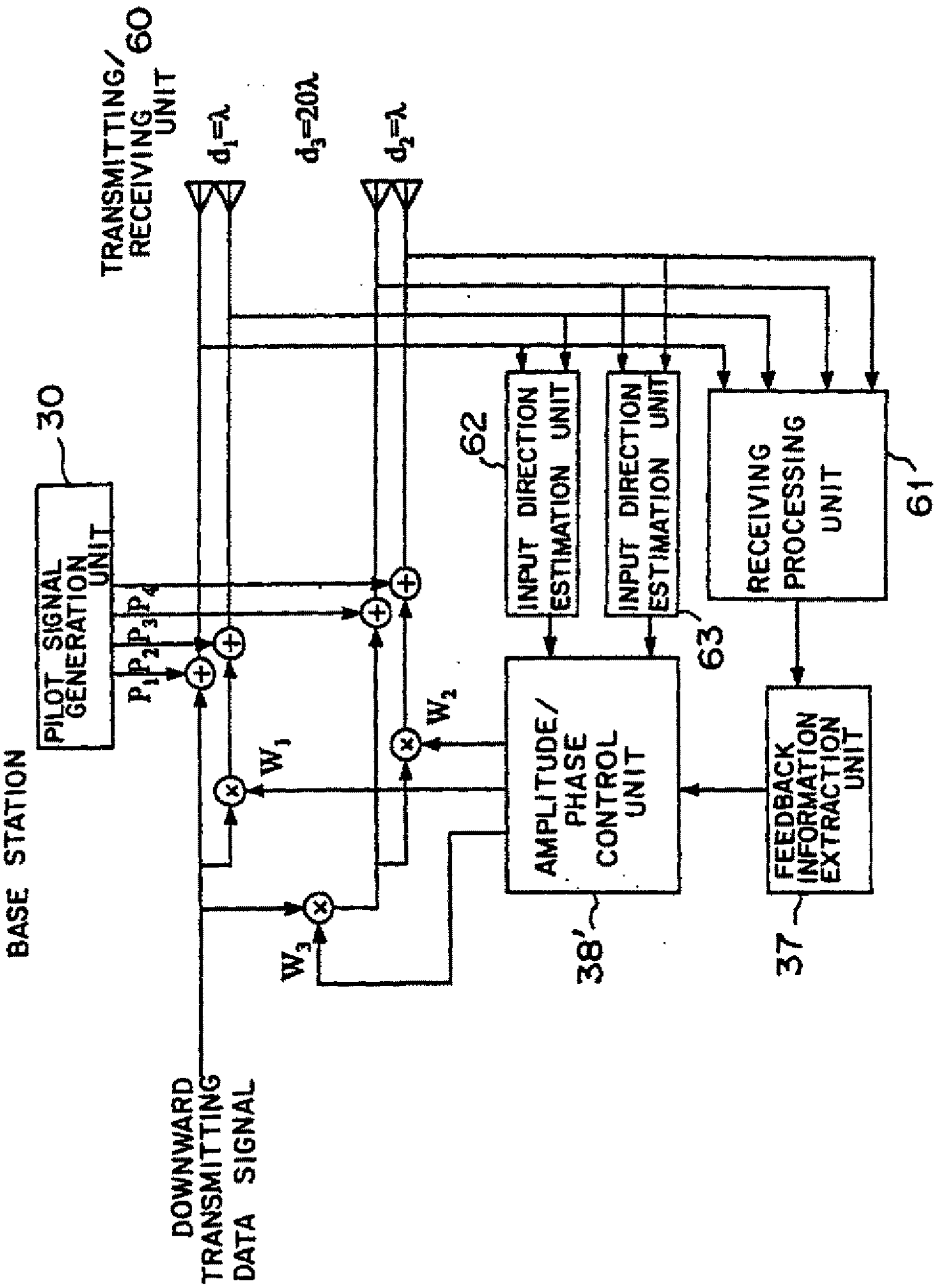

BASE STATION
PILOT SIGNAL GENERATION UNIT
30
P_1 P_2 P_3 P_4
TRANSMITTING/ RECEIVING UNIT
60
$d_1=\lambda$
$d_3=20\lambda$
$d_2=\lambda$
W_1
W_2
W_3
DOWNWARD TRANSMITTING DATA SIGNAL
AMPLITUDE/ PHASE CONTROL UNIT
38'
INPUT DIRECTION ESTIMATION UNIT
62
INPUT DIRECTION ESTIMATION UNIT
63
RECEIVING PROCESSING UNIT
61
FEEDBACK INFORMATION EXTRACTION UNIT
37

FIG. 13

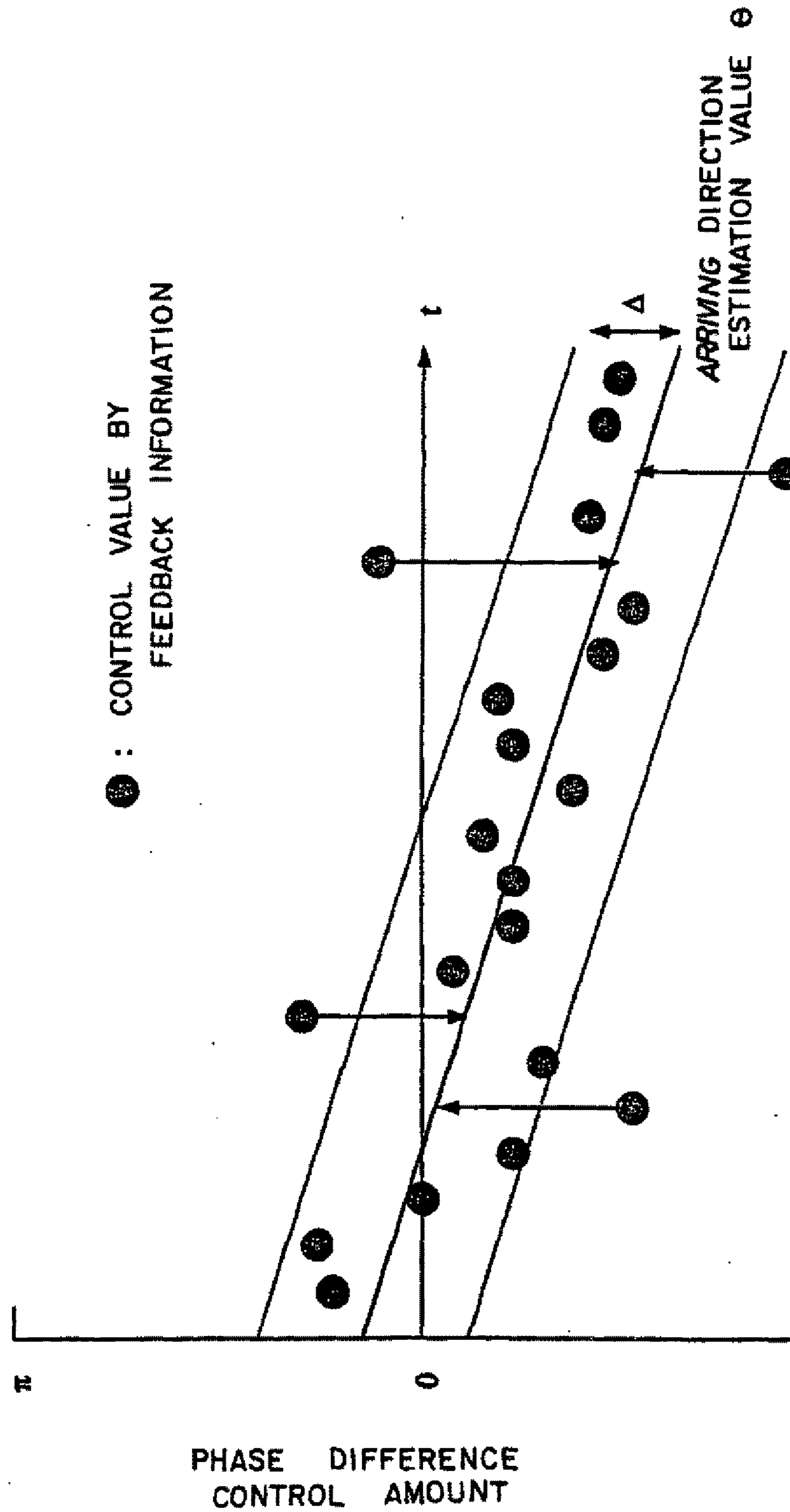

● : CONTROL VALUE BY
FEEDBACK INFORMATION
t
Δ
ARRIVING DIRECTION
ESTIMATION VALUE θ
π
0
-π
PHASE DIFFERENCE
CONTROL AMOUNT

FIG. 14

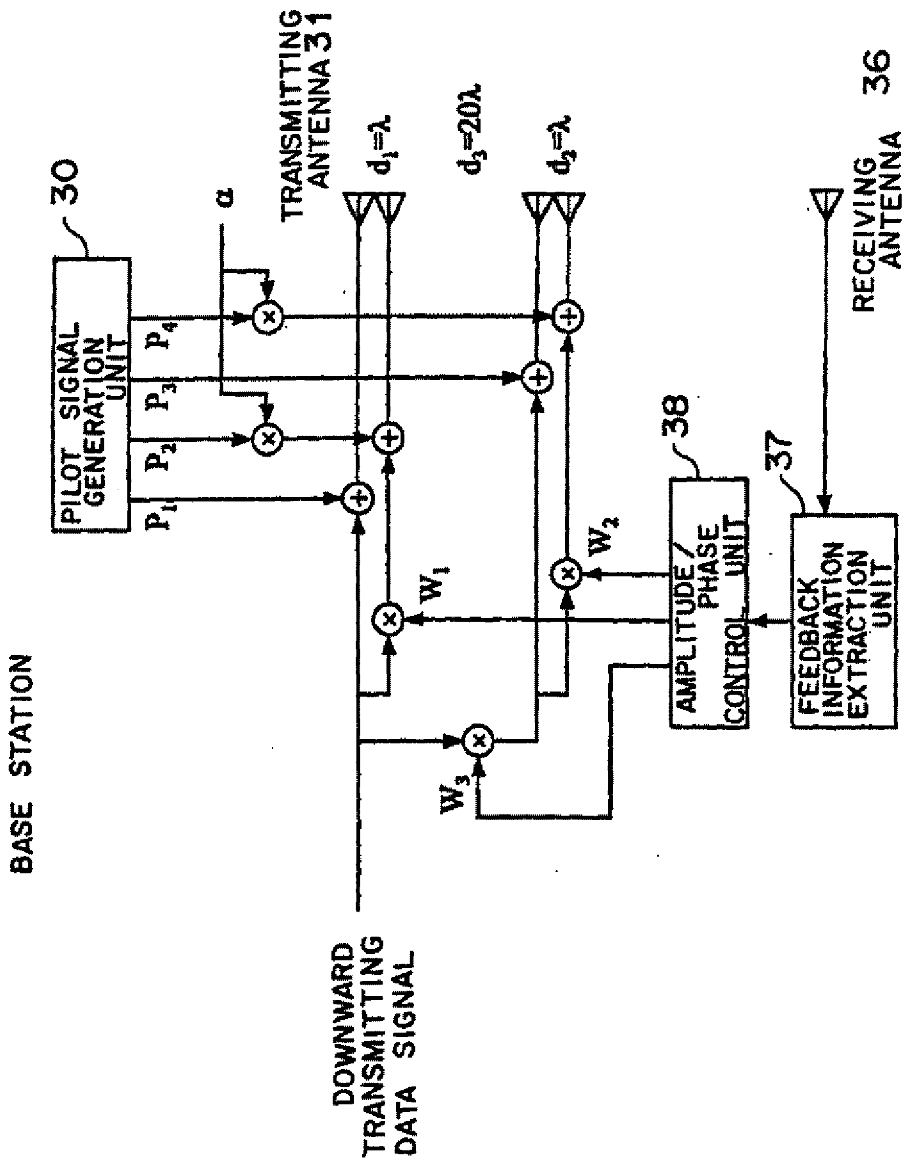

BASE STATION
PILOT SIGNAL GENERATION UNIT
30
P1
P2
P3
P4
α
TRANSMITTING ANTENNA 31
d1=λ
d3=20λ
d2=λ
RECEIVING ANTENNA
36
AMPLITUDE/PHASE CONTROL UNIT
38
FEEDBACK INFORMATION EXTRACTION UNIT
37
W1
W2
W3
DOWNWARD TRANSMITTING DATA SIGNAL

FIG. 15

INTERNATIONAL SEARCH REPORT

International application No.
PCT/JP00/05380

A. CLASSIFICATION OF SUBJECT MATTER
Int.Cl^7 H04B 7/06, 7/10, 7/26, H01Q 3/24

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
Int.Cl^7 H01Q 3/00- 3/46, 21/00-25/04
H04B 7/00, 7/02-7/12, 7/24-7/26, 113
H04L 1/02- 1/06, H04Q7/00-7/04

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
Jitsuyo Shinan Koho 1922-1996 Toroku Jitsuyo Shinan Koho 1994-2000
Kokai Jitsuyo Shinan Koho 1971-2000 Jitsuyo Shinan Toroku Koho 1996-2000

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y A	JP 58-87928 A (Nippon Telegr. & Teleph. Corp. <NTT>), 25 May, 1983 (25.05.83) (Family: none)	20,25 1-19,21-24
A	JP 10-190537 A (NEC Corporation), 21 July, 1998 (21.07.98) (Family: none)	1-25
A	JP 9-200115 A (Toshiba Corporation), 31 July, 1997 (31.07.97) (Family: none)	1-25

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier document but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
23 October, 2000 (23.10.00)	31 October, 2000 (31.10.00)
Name and mailing address of the ISA/ Japanese Patent Office	Authorized officer
Facsimile No.	Telephone No.

Form PCT/ISA/210 (second sheet) (July 1992)

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European Patent Office
Office européen des brevets

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(43) Date of publication:
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(21) Application number: 02254685.7

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AL LT LV MK RO SI

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80336 München (DE)

(54) A wireless communication apparatus and method

(57) A method and apparatus for achieving combined beamforming and transmit diversity for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with at least a single receive antenna, the method comprising the steps of: providing a signal to be transmitted; space-time encoding the signal to produce at least two separate signals, each on a respective output; feeding each output signal to a multiple access transmit processor to produce an output signal; applying respective selected transmit beamforming weights to each output signal; feeding the respective weighted signals to a signal combiner to perform a summing function of the signals and produce a signal for transmission; feeding the summed signal to each of the multiple transmit antennae for transmission; transmitting the signals over respective physical channels; receiving the transmitted signal at at least a single receive antenna; feeding the transmitted signal to a multiple access receive processor to produce an output signal; and space-time decoding the received signal.

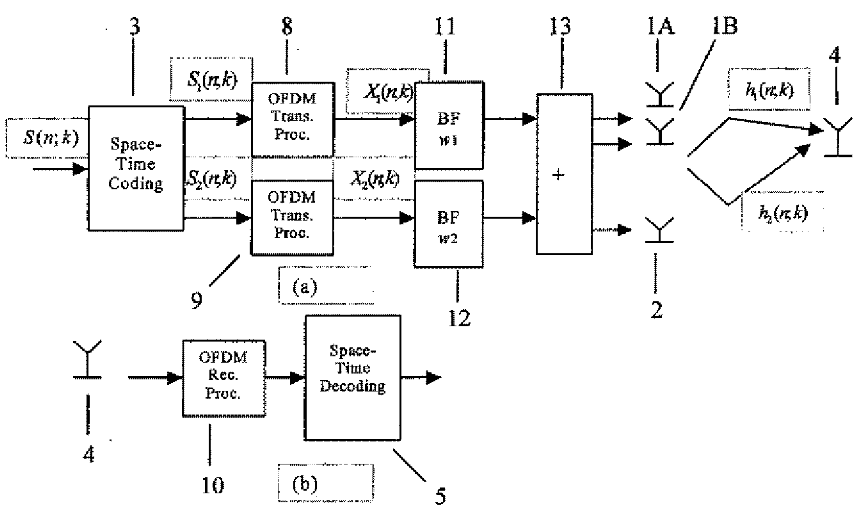


Figure5

Printed by Jouve, 75001 PARIS (FR)

Description

BACKGROUND OF THE INVENTION

[0001] The present invention relates in general to wireless communication systems and, more particularly, to improving the downlink performance of wireless communication systems.

[0002] Wireless mobile communications suffer from four major impairments: path loss, multipath fading, inter-symbol interference (ISI) and co-channel interference. Adaptive antennas can be used to suppress the effects of these factors to improve the performance of wireless communication systems. There are two types of adaptive antennas: diversity antennas and beamforming antennas. In a diversity antenna system, multiple low-correlation or independent fading channels are acquired in order to compensate multipath fading, thus achieving diversity gain. Beamforming antennas, on the other hand, provide beamforming gain by making use of spatial directivity, thus compensating for path loss to a certain extent and suppressing co-channel interference.

[0003] In a diversity antenna system, the antenna spacing is usually required to be large enough, e.g., 10λ in order to obtain low-correlation/independent fading channels, especially for small angular spread environments. However, beamforming antennas need to achieve spatial directivity, so the signals received at and/or transmitted from all antennas must be correlated. This means that for beamforming antenna, the antenna spacing should usually be small, e.g. half wavelength for a uniform linear array (ULA). Because of the conflict between the required antenna spacings for diversity antenna systems and beamforming systems, a prejudice exists that diversity gain and beamforming gain cannot be achieved simultaneously.

SUMMARY OF THE INVENTION

[0004] It is an object of the present invention to seek to provide a wireless communication system benefiting simultaneously from both diversity gain and beamforming gain.

[0005] Accordingly, one aspect of the present invention provides a method of achieving transmit diversity gain in a communication system having a base station with multiple transmit antennae and a mobile terminal with a single receive antenna, the method comprising the steps of: providing a signal to be transmitted $s(n)$; space-time encoding the signal $s(n)$ to produce at least two separate signals $s_1(n)$,$s_2(n)$, each on a respective output; feeding each output signal $s_1(n)$,$s_2(n)$ to a zero-forcing pre-equaliser having a respective function $g_1(k)$, $g_2(k)$ to produce an output signal $x_1(n)$, $x_2(n)$; feeding the output signal $x_1(n)$, $x_2(n)$ of each pre-equaliser to a transmit antenna; transmitting the output signals $x_1(n)$, $x_2(n)$ over respective physical channels $h_1(k)$, $h_2(k)$; receiving the output signals $x_1(n)$, $x_2(n)$ at a single receive antenna; and space-time decoding the received signals, wherein the functions $g_1(k)$, $g_2(k)$ of the zero-forcing pre-equalisers are selected such that the channel responses $g_1(k)^*h_1(k)$, $g_2(k)^*h_2(k)$ of the respective physical channels $h_1(k)$, $h_2(k)$ are flat fading channels.

[0006] Preferably, the communications system is a time-division duplex system and the method includes the further step of deriving the real channel coefficients from uplink channel coefficients for use in selecting the functions $g_1(k)$, $g_2(k)$ of the pre-equalisers.

[0007] Conveniently, the step of deriving the real channel coefficients from uplink channel coefficients uses training symbols from the uplink channel.

[0008] Advantageously, the step of deriving the real channel coefficients from uplink channel coefficients uses blind techniques.

[0009] Preferably, the communications system is a frequency-division duplex system and the method includes the further step of deriving the real channel coefficients by sending a set of training symbols to the receive antenna of the mobile terminal, the mobile terminal estimating the real channel coefficients and feeding back channel coefficient information to the base station.

[0010] Another aspect of the present invention provides a base station with multiple transmit antennae for communicating with a mobile terminal having a single receive antenna over physical channels $h_1(k)$, $h_2(k)$, the base station comprising:

> a space-time encoder having an input of a signal to be transmitted $s(n)$ and at least two outputs each producing a separate signal $s_1(n)$,$s_2(n)$; at least two zero-forcing pre-equalisers, each fed by a respective output signal $s_1(n)$,$s_2(n)$ and having a respective function $g_1(k)$, $g_2(k)$ to produce an output signal $x_1(n)$, $x_2(n)$; and at least two transmit antennae, each being fed by the output signal $x_1(n)$, $x_2(n)$ of a respective one of the pre-equalisers, wherein the functions $g_1(k)$, $g_2(k)$ of the zero-forcing pre-equalisers are selected such that the channel responses $g_1(k)^*h_1(k)$, $g_2(k)^*h_2(k)$ of the respective physical channels $h_1(k)$, $h_2(k)$ are flat fading channels.

[0011] Preferably, the mobile terminal has a single receive antenna and a space-time decoder to decode the signals

received from the base station.

[0012] A further aspect of the present invention provides a method of achieving combined beamforming and transmit diversity for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with a single receive antenna, the method comprising the steps of: providing a signal to be transmitted $S(n;k)$; space-time encoding the signal $S(n;k)$ to produce at least two separate signals $S_1(n;k), S_2(n;k)$, each on a respective output; feeding each output signal $S_1(n;k), S_2(n;k)$ to a transmit processor to produce an output signal $X_1(n;k)$, $X_2(n;k)$; applying respective selected transmit beamforming weights to each output signal $X_1(n;k)$, $X_2(n;k)$; feeding the respective weighted signals to a signal combiner to perform a summing function of the signals and produce a signal X(n;k) for transmission; feeding the summed signal X(n;k) to each of the multiple transmit antennae for transmission; transmitting the signals X(n;k) over respective the physical channel h(n;k); receiving the received signal Y(n;k) at a single receive antenna; feeding the received signal Y(n;k) to a receive processor to produce an output signal; and space-time decoding the received signal.

[0013] Preferably, the respective transmit beamforming weights are selected as the eigenvectors corresponding to the two largest eigenvalues of the downlink channel covariance matrix (DCCM) of the physical channel h(n;k).

[0014] Conveniently, the physical channel h(n;k) consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, and the transmit processors do not add cyclic prefixes and one of the output signals from the transmit processors is delayed by $\Delta\tau$ before the respective selected transmit beamforming weight is applied thereto, the beamforming weights being chosen such that the delayed signal or its inverse fast Fourier transform (IFFT) only goes through one channel $h_1(n;k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal or its IFFT only goes through another channel $h_2(n;k)$ between the base station multiple transmit antennae and the receive antenna, thereby creating two different channels which can be space-time decoded to recover the transmitted signal.

[0015] Advantageously, the physical channel h(n;k) consists of two time-delayed clustered rays, $h_1(n;k)$ and $h_2(n;k)$, the transmit processors have a cyclic prefix length of $\Delta\psi$ and one of the output signals from the transmit processors is delayed by ψ before the respective selected transmit beamforming weight is applied thereto, the beamforming weights being chosen such that the delayed signal or its inverse fast Fourier transform (IFFT) only goes through one channel $h_1(n;k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal or its IFFT only goes through another channel $h_2(n;k)$ between the base station multiple transmit antennae and the receive antenna, thereby creating two different channels which can be space-time decoded to recover the transmitted signal.

[0016] Preferably, the method comprises the further steps of: estimating a power-delay-DOA profile for the channel h(n;k); and, based on the profile: determining the cyclic prefix length, $\Delta\psi$, to be added by the transmit processors; determining the delay ψ; and determining the transmit beamforming weights.

[0017] Advantageously, the method comprises the further step of estimating the downlink channel covariance matrix (DCCM) from the uplink channel covariance matrix (UCCM) to construct transmit beamforming weights.

[0018] Conveniently, the method comprises the further steps of: estimating the downlink channel covariance matrix (DCCM) from the uplink channel covariance matrix (UCCM) to construct transmit beamforming weights; estimating a power-delay-DOA profile for channel h(n;k); and, based on the profile: determining the length, $\Delta\psi$, of the cyclic prefix to be added by the transmit processors; determining the delay ψ; and determining the transmit beamforming weights.

[0019] A further aspect of the present invention provides a base station with multiple transmit antennae for communicating with a mobile terminal having a single receive antenna over physical channel h(n;k) having two time-delayed rays, h_1(n;k) and h_2(n;k), the base station comprising:

a space-time encoder having an input of a signal to be transmitted and at least two outputs each producing a separate signal; at least two transmit processors each receiving one of the outputs from a respective space-time encoder; at least two transmit beamformers each receiving an output from a respective transmit processor and applying a transmit beamforming weight thereto; a signal combiner receiving signals from the beamformers and operable to perform a summing function of the signals from the beamformers and produce a signal for transmission by the multiple transmit antennae.

[0020] Preferably, a delay of $\Delta\tau$ is interposed between one of the transmit processor outputs and a beamformer to delay the signal output from the transmit processor by $\Delta\tau$ before the respective selected transmit beamforming weight is applied thereto, wherein the transmit processors do not add cyclic prefixes.

[0021] Conveniently, a delay of ψ is interposed between one of thetransmit processor outputs and a beamformer to delay the signal output from the transmit processor by ψ before the respective selected transmit beamforming weight is applied thereto, the transmit processors having a cyclic prefix length of $\Delta\psi$.

[0022] Advantageously, a processor to determine a power-delay-DOA profile estimate for channel h(n;k) is provided and, based on the profile, determine: the length, $\Delta\psi$ cyclic prefix to be added by the transmit processors; the delay ψ; and the transmit beamforming weights.

[0023] Conveniently, a processor is provided to estimate a downlink channel covariance matrix (DCCM) from the

uplink channel covariance matrix (UCCM) to construct transmit beamforming weights.

[0024] Preferably, the base station further comprises a first processor to determine a power-delay-DOA profile estimate for channel $h(n;k)$; and, based on the profile, determine: the length, $\Delta\psi$, of the cyclic prefix to be added by the transmit processors; the delay ψ; and the transmit beamforming weights; and a second processor to estimate a downlink channel covariance matrix (DCCM) from the uplink channel covariance matrix (UCCM) to construct transmit beamforming weights.

[0025] Conveniently, the transmit and receive processors are selected from the group consisting of: OFDM, CDMA and TDMA processors.

[0026] Advantageously, the communications system comprises the base station and a mobile terminal having a single receive antenna, a receive processor to produce an output signal and a space-time decoder to decode the output signal.

[0027] A further aspect of the present invention provides a method of achieving combined beamforming and transmit diversity for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with a single receive antenna, the method comprising the steps of: providing a signal to be transmitted $s(n)$; space-time encoding a signal to be transmitted $s(n)$ to produce at least two separate signals $s_1(n), s_2(n)$, each on a respective output; delaying one of the space-time encoded output signals by $\Delta\tau$; applying respective selected transmit beamforming weights to the delayed and undelayed signals; feeding the respective weighted signals to a signal combiner to perform a summing function of the signals and produce a signal for transmission; feeding the summed signal to each of the multiple transmit antennae for transmission; transmitting the summed signals over the physical channel $h(k)$ with two time-delayed rays $h_1(k)$, $h_2(k)$; receiving the major components of the transmitted signals at a single receive antenna at substantially the same time; and space-time decoding the received signal.

[0028] Preferably, the beamforming weights are chosen such that the delayed signal only goes through one ray $h_1(k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal only goes through another ray $h_2(k)$ between the base station multiple transmit antennae and the receive antenna.

[0029] Conveniently, the delay $\Delta\tau$ is derived from downlink channel information.

[0030] A further aspect of the present invention provides a base station with multiple transmit antennae for communicating with a mobile terminal having a single receive antenna over physical channel $h(k)$ having two time-delayed rays $h_1(k)$, $h_2(k)$, the base station comprising:

a space-time encoder having an input of a signal to be transmitted and at least two outputs each producing a separate signal; at least two transmit beamformers each receiving an output from the space-time encoder and applying a transmit beamforming weight thereto; a signal combiner receiving signals from the beamformers and operable to perform a summing function of the signals from the beamformers and produce a signal for transmission by each of the multiple transmit antennae, wherein a delay of $\Delta\tau$ is interposed between the space-time encoder and one of the beamformers such that the major components of the transmitted signals are received at a single receive antenna at substantially the same time.

[0031] Preferably, the communications system comprises the base station and a mobile terminal having a single receive antenna and a space-time decoder to decode the received signal.

[0032] One aim of the present invention is to seek to achieve, at the mobile terminal, diversity gain, beamforming gain as well as delay spread reduction simultaneously by using a base station with a multiple antenna array.

[0033] The advantages of the embodiments of the present invention are as follows:

- Beamforming gain and transmit diversity are achieved simultaneously;
- Based on power-delay-DOA profile, delay spread is reduced adaptively.
  - In two-ray environment, a frequency selective fading channel is transferred into a flat fading channel, yet the path diversity gain is maintained.
  - In hilly terrain (HT) environment, we can transfer a long delay spread channel into a short delay spread channel, yet still maintain the path diversity gain.
- With delay spread reduction and combined beamforming and transmit diversity, the invented systems provide high spectrum efficiency, yet consumes less transmission power.
- The invented systems also employ adaptive modulation to further improve the spectrum efficiency based on the diversity order and channel conditions.
- The mobile terminal is usually limited by physical size and battery power. The invented systems put the complicated processing at the base station, rather at the mobile terminal. Thus the mobile terminal complexity is reduced.
- The invented systems are well applicable for the applications which require high data rate for downlink transmission.

These applications include, for example, high speed downlink packet access (HSDPA) in 3rd generation partnership project (3GPP), wireless internet, and wireless multimedia communications.

[0034] In order that the present invention may be more readily understood, embodiments thereof will now be described, by way of example, with reference to the accompanying drawings, in which:

Figure 1 (*Prior Art*) is a schematic diagram illustrating Alamouti's permutation transmit diversity method;

Figure 2 is a schematic diagram illustrating a method embodying the present invention using transmit diversity with pre-equalization for frequency selective fading channels;

Figure 3 *(Prior Art)* is a schematic diagram illustrating orthogonal frequency division multiplexing (OFDM) with transmit diversity at: (a) a transmitter; and (b) a receiver;

Figure 4 *(Prior Art)* is a schematic diagram illustrating OFDM combined beamforming and transmit diversity for flat fading channels;

Figure 5 is a schematic diagram illustrating a method embodying the present invention using OFDM with combined beamforming and transmit diversity at: (a) a transmitter; and (b) a receiver;

Figure 6 is a schematic diagram illustrating a method embodying the present invention using combined beamforming and transmit diversity for two ray (TR) frequency selective fading channels at (a) a transmitter; and (b) a receiver;

Figure 7 is a schematic diagram illustrating a method embodying the present invention using OFDM with combined beamforming and transmit diversity for two ray (TR) models at: (a) a transmitter; and (b) a receiver;

Figure 8 is a schematic diagram illustrating a method embodying the present invention using OFDM with combined beamforming and transmit diversity for hilly-terrain (HR) models at (a) a transmitter; and (b) a receiver; and

Figure 9 is a schematic diagram illustrating a method embodying the present invention using OFDM with combined beamforming, transmit diversity and adaptive delay spread reduction: at (a) a transmitter; and (b) a receiver.

DETAILED DESCRIPTION OF THE INVENTION

[0035] The present invention revolves around the use of multiple antennas at the base station to improve the downlink performance of a wireless communication system. Downlink beamforming is effective in limiting interference pollution, which is of critical importance especially in multimedia communications. Transmit diversity is a powerful technique when receive diversity is impractical, especially for mobile terminals with size and/or power limitations. It can also be used to further improve downlink performance even though receive diversity is available.

[0036] In a multipath propagation environment, a receiver acquires several time-delayed, amplitude-scaled and direction of arrival (DOA) dependent versions of a transmitted signal. When the maximum time delay between the first-arrived and last-arrived versions of a signal along the various paths is smaller than the symbol interval, these paths are not resolvable in the time domain. However, these paths are resolvable in the spatial domain as they may come from different DOAs. Since each path may experience independent fading, using a beamforming antenna array, one obtains several independent channels, to which transmit diversity is applicable.

[0037] When the maximum relative delay is greater than the symbol interval, a frequency selective fading channel is observed. Frequency selectivity is beneficial for achieving diversity, however, it also yields inter-symbol interference (ISI) which needs to be suppressed at the receiver. This phenomenon becomes more and more prevalent as the data transmission rate increases. One way to suppress ISI is to use equalization at the receiver. The performance of an equalizer, however, depends on the frequency responses of the wireless channels. Specifically, when the channel's frequency responses have deep nulls in a certain frequency band, the equalization output yields noise enhancement, the effect of which can degrade the diversity gain obtained by the frequency selectivity. On the other hand, An adaptive equalizer often promotes error propagation problems when decision-directed symbols are used as reference signals, and the complexity of the equalizer is further complicated if the delay spread is large.

[0038] Another method of reducing ISI is to reduce the delay spread using adaptive antennas at the base station. For example, if the base station knows the direction-of-arrival (DOA) information of each delayed version of the received signal, it can then form a beam to one path whilst arranging for nulls or small antenna gains at the DOAs of the other paths. In this manner, the mobile terminal only receives one path of each transmitted signal. This method, though

simple in signal detection, sacrifices the diversity gain since use is only being made of one path.

[0039] Compared to receive diversity, transmit diversity has received greater attention during the past decade. Delay diversity as disclosed in A. Wittneben, "A new bandwidth efficient transmit antenna modulation diversity scheme for linear digital modulation", Proc. Of ICC'93, pp. 1630-1634, 1993, is one early transmit diversity technique using multiple transmit antennas. This method transforms a flat fading channel into a frequency selective fading channel making use of frequency diversity. An equalizer is provided at the mobile terminal in order to compensate for the artificially induced ISI. The performance of the equalizer depends on the frequency property of the channels. Further, an adaptive equalizer often promotes error propagation problems when decision-directed symbols are used as reference signals. In fact, it is shown in Y.C. Liang, Y. Li and K.J.R. Liu, "Feasibility of transmit diversity for IS-136 TDMA systems", Proc. Of VTC '98, pp. 2321-2324, 1998, that when the maximum Doppler frequency is over 40 Hz, this diversity method is even worse than that without diversity. In S.M. Alamouti, "A simple transmit diversity technique for wireless communications", IEEE Journal of Selected Areas in Communications, Vol.16, No.8, pp.1451-1458, October 1998, Alamouti proposed a permutation diversity method, whose performance is similar to maximal-ratio combining (MRC) receive diversity. This method only requires a simple receiver structure. More general transmit diversity methods are referred to as space-time coding methods as disclosed in V. Tarokh, N. Seshadri and A.R. Calderbank, "Space-time codes for high data rate wireless communication: Performance analysis and code construction", IEEE trans. On Information Theory, vol. 44, No. 3, pp. 744-765, March 1998. Space-time codes include space-time trellis codes (STTC) and space-time block codes (STBC). In fact, permutation diversity is the simplest class of STBC.

[0040] Figure 1 illustrating Alamouti's permutation diversity method shows the permutation diversity method with two transmit antennas 1, 2 equipped at the base station (BS). The signal $s(n)$ to be transmitted is first coded in a space-time coding module 3 The space-time coding module 3 works in the following way. It has one input port and two output ports. The input port accepts the transmitted sequence, s(0), s(1), ... The two output ports provide, in response, respective output signals $s_1(t)$ and $s_2(t)$ at time instants $t=n$ and $t=n+1$, where n is an even integer, as follows.

	$t=n$	$t=n+1$
$s_1(t)$	$s(n)/\sqrt{2}$	$s^*(n+1)/\sqrt{2}$
$s_2(t)$	$s(n+1)/\sqrt{2}$	$-s^*(n)/\sqrt{2}$

[0041] At a single receive antenna 4 at the mobile terminal the signals received at time instants $t=n$ and $t=n+1$ are given by

$$x(n) = \alpha_1 s_1(n) + \alpha_2 s_2(n) + w(n) \tag{1}$$

$$x(n+1) = \alpha_1 s_1(n+1)+\alpha_2 s_2(n+1)+w(n+1) \tag{2}$$

where α_1 and α_2 are the respective channel responses from the two transmit antennas 1, 2 to the receiver antenna 4, respectively; $w(n)$ is additive white Gaussian noise (AWGN).

[0042] The received signal is subsequently decoded by the space-time decoding module as follows. Specifically, equations (1) and (2) can be written in matrix forms:

$$\begin{bmatrix} x(n) \\ x(n+1) \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} s(n) & s(n+1) \\ s^*(n+1) & -s^*(n) \end{bmatrix} \begin{bmatrix} \alpha_1 \\ \alpha_2 \end{bmatrix} + \begin{bmatrix} w(n) \\ w(n+1) \end{bmatrix} \tag{3}$$

$$\begin{bmatrix} x(n) \\ x^*(n+1) \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} \alpha_1 & \alpha_2 \\ -\alpha_2^* & \alpha_1^* \end{bmatrix} \begin{bmatrix} s(n) \\ s(n+1) \end{bmatrix} + \begin{bmatrix} w(n) \\ w^*(n+1) \end{bmatrix} \tag{4}$$

[0043] Therefore, channel coefficients can be estimated via equation (3) using training symbols; while equation (4) can be used for signal estimation/detection. This signal detection method is also called permutation decoding.

[0044] It is pointed out that, as opposed to delay diversity techniques which require a complicated equalizer at the receiver, the channel estimation and signal detection for permutation diversity involves very simple numerical operations. Also, compared to a one-transmitter/two-receiver receive diversity technique, even though the permutation diversity method has a 3 dB performance loss, it achieves the same order of diversity gain as receive diversity techniques using a maximal ratio combining (MRC) approach.

[0045] Permutation diversity can be extended to space-time block codes (STBC) and space-time trellis codes (STTC). All these codes achieve transmit diversity for flat fading environment.

[0046] One example of the invention applies Alamouti's diversity method to frequency selective fading channels. When the delay spread is greater than the symbol interval, frequency selective fading channels are observed. Figure 2 illustrates the system model applying Alamouti's diversity method to frequency selective fading channels. The transmitted signal, $s(n)$, is first coded using Alamouti's codes in the coding module 3, with the two branch outputs as $s_1(n)$ and $s_2(n)$. $s_1(n)$ and $s_2(n)$ are then passed into two pre-equalizers, 6, 7 having functions $g_1(k)$ and $g_2(k)$, to produce two output sequences $y_1(n)$ and $y_2(n)$. $y_1(n)$ and $y_2(n)$ are finally modulated and up-converted as RF signals, which are sent out through the transmit antennas 1, 2 as physical channels $h_1(k)$ and $h_2(k)$.

[0047] The functions $g_1(k)$ and $g_2(k)$ of the pre-equalizers 6,7 are used to pre-equalize the two physical channels, $h_1(k)$ and $h_2(k)$, respectively. By designing the pre-equalizers with zero-forcing criterion, the overall channel responses, $g_1(k) * h_1(k)$ and $g_2(k) * h_2(k)$, are now flat fading channels, with which Alamouti's coding/decoding method can be used. Here, " * " denotes a convolution operation.

[0048] In order to design the pre-equalizers 6,7, the real channel coefficients, $h_1(k)$ and $h_2(k)$, should be known at the base station/transmit antennas 1, 2. This can be done in two ways. For time-division duplex (TDD) systems, downlink channel coefficients are the same as uplink channel coefficients, which are derivable from the uplink using training symbols or blind techniques (up to a constant scaler). For frequency-division duplex (FDD) systems, the base station sends a set of training symbols to the mobile terminal, which then estimates and feeds back the downlink channel information to the base station.

[0049] The above methods are also applicable for other space-time codes.

[0050] Orthogonal frequency division multiplexing (OFDM) is a known and effective method of combatting the large delay spread problem. The combination of OFDM with a transmit diversity method not only suppresses large delay spread, but also achieves transmit diversity gain. Figure 3 shows a prior art OFDM system with two-antenna transmit diversity as described in Y. Li, N. Seshadri and S. Ariyavisitakul, "Channel estimation for OFDM systems with transmitter diversity in mobile wireless channels", IEEE Journal of Selected Areas in Communications, vol. 17, No. 3, pp. 461-471, March 1999. The signal to be transmitted, $S(n;k)$, is first coded using space-time codes in coding module 3, yielding two branch outputs as $S_1(n;k)$ and $S_2(n;k)$. $S_1(n;k)$ and $S_2(n;k)$ are then passed into respective normal OFDM transmit processors 8, 9, whose outputs are finally modulated and up-converted as RF signals, which are sent out through transmit antennas 1, 2.

[0051] At the single antenna receiver 4 at the mobile station, the received signal is passed into a normal OFDM receive processor 10, followed by a space-time decoder module 5. Specifically, the fast Fourier transform (FFT) output becomes

$$X(n;k) = H_1(n;k)S_1(n;k) + H_2(n;k)S_2(n;k) + W(n;k) \tag{5}$$

$$X(n;k+1) = H_1(n;k+1)S_1(n;k+1) + H_2(n;k+1)S_2(n;k+1) + W(n;k+1) \tag{6}$$

[0052] In (5) and (6), $H_1(n;k)$ and $H_2(n;k)$ are, respectively, the Fourier transforms of the channel impulse responses, $h_1(n;k)$ between transmit antenna 1 and receive antenna 4, and $h_2(n;k)$ between transmit antenna 2 and receive antenna 4; $W(n;k)$ is the FFT output of the additive noise, $w(n;k)$, received at the receive antenna 4.

[0053] Permutation decoding methods can be easily applied if $S_1(n;t)$ and $S_2(n;t)$ at time instants $t=k$ and $t=k+1$, where k is an even integer, are chosen as follows:

	$t=k$	$t=k+1$
$S_1(n;t)$	$S(n;k)\sqrt{2}$	$S^*(n;k+1)/\sqrt{2}$
$S_2(n;t)$	$S(n;k+1)/\sqrt{2}$	$-S^*(n;k)/\sqrt{2}$

Prior Art: Combined beamforming and transmit diversity for flat fading channels.

[0054] The above three methods (Alamouti's permutation diversity method, a diversity method applied to frequency selective fading channels and OFDM with transmit diversity) achieve transmit diversity gain for flat fading channels, or frequency selective fading channels. The transmit antennas belong to diversity antennas, i.e., the antenna spacing is large, e.g., ten times wavelength, typically.

[0055] Figure 4 shows a known system combining beamforming and transmit diversity for flat fading channels as disclosed in R. Negi, A.M. Tehrani and J. Cioffi, "Adaptive antennas for space-time coding over block invariant multipath fading channels", Proc. of IEEE VTC, pp. 70-74, 1999. The signal to be transmitted, $s(n)$, is first coded using a space-time coder module 3, yielding two branch outputs as $s_1(n)$ and $s_2(n)$. $s_1(n)$ and $s_2(n)$ are then passed into two transmit beamformers 11,12, w_1 and w_2, respectively, followed by a signal combiner 13 which performs a simple summing function of the two inputs to producing a signal $x(n)$ for transmission which, in vector form, is as follows:

$$x(n) = w_1^H\, s_1(n) + w_2^H\, s_2(n) \tag{7}$$

[0056] To obtain spatial selectivity, the antenna spacing, d, is set to be small, e.g., half wavelength, and the number of transmit antennas 1A, 1B, 2, M, is greater than two. This is a beamforming antenna array, instead of a diversity antenna array. Suppose the physical channel consists of L spatially separated paths, whose fading coefficients and DOAs are denoted as $(\alpha_k(t),\theta_k)$, for $k = 1,\cdots,L$. If the maximum time delay relative to the first arrived path is smaller than the symbol interval, a flat fading channel is observed, and the instantaneous channel response, $h_d(t)$, can be expressed as follows:

$$h_d(t) = \sum_{k=1}^{L} \alpha_k(t) a_d(\theta_k) \tag{8}$$

where $a_d(\theta_k)$ is the downlink steering vector at DOA θ_k. The received signal, $y(n)$, at the mobile terminal is given by

$$y(n) = w_1^H h_d(t) s_1(n) + w_2^H h_d(t) s_2(n) + w(n) \tag{9}$$

[0057] By denoting $\beta_1(t)=w_1^H h_d(t)$, $\beta_2(t)=w_2^H h_d(t)$, the transmit beamforming weights can be estimated by maximizing the cost function:

$$J = E|\beta_1(t)|^2 + E|\beta_2(t)|^2 \tag{10}$$

$$\text{s.t.}\quad E[\beta_1(t)\beta_2^*(t)] = 0 \tag{11}$$

[0058] Maximum average signal to noise ratio (SNR) is obtained by maximising (10); while condition (11) guarantees that $\beta_1(t)$ and $\beta_2(t)$ are statistically uncorrelated, thus maximum diversity gain can be achieved.

[0059] Comparing (9) with (1), with the aid of downlink beamforming, two statistical uncorrelated fading channels, $\beta_1(t)$ and $\beta_2(t)$ have been artificially generated, with which space-time decoding can be used to recover the transmitted signal, $s(n)$. For Alamouti's diversity method, permutation decoding is applied.

[0060] The optimal transmit beamforming weight vectors are the eigenvectors corresponding to the two largest eigenvalues of the downlink channel covariance matrix (DCCM):

$$R_d = E\left[h_d(t)h_d^H(t)\right] \quad (12)$$

where the expectation is conducted over all fading coefficients. Suppose all paths have the same average power, or $E|\alpha_k(t)|^2 = 1/L$, the DCCM is given by

$$R_d = \frac{1}{L}\sum_{k=1}^{L} a_d(\theta_k)a_d^H(\theta_k) \quad (13)$$

[0061] For TDD, DCCM is the same as uplink channel covariance matrix (UCCM). For FDD, there are two ways to estimate the DCCM, both of which are based on the fact that uplink and downlink signals go through the same DOAs. The first method estimates the DOAs of all paths from the received uplink signals first, then constructs the downlink steering vectors, $a_d(\theta_k)$'s, and further DCCM R_d via equation (13). The second method estimates DCCM from UCCM directly via frequency calibration processing as disclosed in Y-C. Liang and F. Chin, "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4. This method does not involve DOA estimation and its associates and is therefore simple to implement.

[0062] This system achieves diversity gain and beamforming gain simultaneously for flat fading environment but it is desirable to extend that system into a frequency selective fading environment.

[0063] For mobile wireless communications without beamforming, the two ray (TR) model, typical urban (TU) model, and hilly terrain (HT) model are three commonly used power-delay profiles. When downlink beamforming is added, a power-delay-DOA profile should be considered. In picocell, microcell, and macrocell with TU model, there is less correlation between path delays and the DOAs. However, in macrocell with TR and HR models, the path delays are usually statistically dependent on the DOAs. We will show that for different environments, there exist different schemes to achieve combined beamforming and transmit diversity gains, as well as maximum spectrum efficiency.

[0064] Another example of the invention utilises OFDM to obtain combined beamforming and transmit diversity.

[0065] Combined beamforming and transmit diversity can be achieved by using OFDM for frequency selective fading channels. Figure 5 shows the OFDM system with combined beamforming and transmit diversity. Even though OFDM is selected as one example to show how the delay spread can be reduced, while yet maintaining beamforming and transmit diversity gain, other examples being other multi-carrier modulation schemes, such as MC-CDMA, MC-DS-CDMA and single carrier systems with cyclic prefix.

[0066] The transmitted signal at the kth tone of the nth block, $S(n;k)$, is first coded at the base station using space-time codes in coding module 3, yielding two branch outputs, $S_1(n;k)$ and $S_2(n;k)$. $S_1(n;k)$ and $S_2(n;k)$ are passed into respective normal OFDM transmit processors 8,9, followed by two transmit beamformers, 10,11, (w_1 and w_2) respectively. The beamforming outputs are finally combined in a combiner 13, and transmitted out through the transmit antennas 1A, 1B, 2 of the base station antenna array.

[0067] With the base station antenna array 1A, 1B, 2, the complex baseband representation of a wireless channel impulse response can be described as the following vector form

$$h_d(t;\tau) = \sum_m \sum_l \gamma_{m,l}(t)a_d(\theta_{m,l})\delta(\tau - \tau_m) \quad (14)$$

where τ_m is the delay of the mth path resolved in time, $\gamma_{m,l}(t)$ and $a_d(\theta_{m,l})$ are the complex amplitude and downlink steering vector corresponding to lth DOA of the mth delay path. Because of the motion of the vehicular, $\gamma_{m,l}(t)$'s are wide-sense stationary (WSS) narrow band complex Gaussian processes, which are zero-mean and statistically independent for different m's, or l's. Suppose all $\gamma_{m,l}(t)$'s have the same normalized correlation function, $r(t)$ ($r(0)=1$), but possibly different average power, σ^2_{ml}, then

$$E\left[\gamma_{m,l}(t+\Delta t)\gamma^{*}_{m,l}(t)\right]=\sigma^{2}_{m,l}r(\Delta t) \tag{15}$$

[0068] The Fourier transform (FT) of $h(t;\tau)$ at time instant t is given by

$$H_d(t;f)=\int_{-\infty}^{\infty}h_d(t;\tau)e^{-j2\pi f\tau}d\tau=\sum_m\sum_l\gamma_{m,l}(t)a_d(\theta_{m,l})e^{-j2\pi f\tau_m} \tag{16}$$

[0069] For an OFDM system with block length T_b and tone spacing f_t, the discrete value of $H(t;f)$ is given by

$$H_d[n;k]\underline{\underline{\Delta}}H_d(nT_b;kf_t)=\sum_m\sum_l\gamma_{m,l}(nT_s)a_d(\theta_{m,l})e^{-j2\pi kf_t\tau_m} \tag{17}$$

thus the correlation function matrix of the frequency response for different times and frequencies is given by

$$r_d[\Delta n;\Delta k]=E\left[H_d[n+\Delta n;k+\Delta k]H_d^H[n;k]\right]=r(\Delta nT_b)\sum_m e^{-j2\pi\Delta kf_t\tau_m}R_{d,m} \tag{18}$$

where

$$R_{d,m}=\sum_l\sigma^2_{m,l}a_d(\theta_{m,l})a_d^H(\theta_{m,l})$$

is the downlink channel covariance matrix corresponding to the mth delay path. Note for $\Delta n = 0$ and $\Delta k = 0$,

$$r_d[0;0]=\sum_m\sum_l\sigma^2_{m,l}a_d(\theta_{m,l})a_d^H(\theta_{m,l})\underline{\underline{\Delta}}R_d \tag{19}$$

[0070] At the mobile terminal single antenna 4, the received signals are first passed into normal OFDM receive processor 10, followed by a permutation decoder 5. Within the normal OFDM receive processor, the FFT output becomes

$$X[n;k]=w_1^H H_d[n;k]S_1[n;k]+w_2^H H_d[n;k]S_2[n;k]+W[n;k] \tag{20}$$

$$X[n;k+1] = w_1^H H_d[n;k+1]S_1[n;k+1]+w_2^H H_d[n;k+1]S_2[n;k+1]+W[n;k+1] \quad (21)$$

where $W[n;K]$ is zero mean AWGN.

[0071] By denoting $\beta_1 = w_1^H H_d[n;k]$, $\beta_2 = w_2^H H_d[n;k]$, the beamforming weights can be estimated by maximizing the cost function:

$$J = E|\beta_1|^2 + E|\beta_2|^2 \quad (22)$$

$$\text{s.t. } E[\beta_1\beta_2^*] = 0 \quad (23)$$

[0072] Again, maximum average SNR is obtained through maximizing equation (22); while condition (23) guarantees that β_1 and β_2 are statistically uncorrelated, thus maximum diversity gain can be achieved.

[0073] The optimal transmit beamforming weight vectors are the eigenvectors corresponding to the two largest eigenvalues of downlink channel covariance matrix (DCCM) R_d.

$$R_d = E\left[H_d[n;k]H_d^H[n;k]\right] \quad (24)$$

[0074] Comparing equations (20) and (21) with equations (5) and (6), with the aid of downlink beamforming, two uncorrelated fading channels are generated, with which the space-time decoding can be used to recover the transmitted signal. Permutation decoding method can be applied if $S_1(n;k)$ and $S_2(n;k)$ are chosen as follows.

	$t=k$	$t=k+1$
$S_1(n;t)$	$s(n;k)/\sqrt{2}$	$s^*(n;k+1)/\sqrt{2}$
$S_2(n;t)$	$s(n;k+1)/\sqrt{2}$	$-s^*(n;k)/\sqrt{2}$

[0075] A frequency calibration method for DCCM estimation for OFDM.

[0076] In order to generate the downlink beamforming weights, it is first necessary to construct the DCCM. A frequency calibration (FC) method disclosed in Y-C. Liang and F. Chin, "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4 is applied.

[0077] Using a similar method, we can show that the correlation function matrix of the uplink frequency response for different times and frequencies is given by

$$r_u[\Delta n;\Delta k] = E\left[H_u[n+\Delta n;k+\Delta k]H_u^H[n;k]\right] = r(\Delta nT_b)\sum_m e^{-j2\pi\Delta kf_s\tau_m}R_{u,m} \quad (25)$$

where

$$R_{u,m} = \sum_l \sigma_{m,l}^2 a_u(\theta_{m,l})a_u^H(\theta_{m,l})$$

is the uplink channel covariance matrix corresponding to the mth delay path. Note for $\Delta n = 0$ and $\Delta k = 0$,

$$r_u[0;0] = \sum_m \sum_l \sigma_{m,l}^2 a_u(\theta_{m,l}) a_u^H(\theta_{m,l}) \underline{\underline{\Delta}} R_u \tag{26}$$

[0078] Comparing equations (19) and (26), the FC method devised in Y-C. Liang and F. Chin, "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4 is used to estimate the DCCM from UCCM.

[0079] This system provides diversity gain and beamforming gain for OFDM systems. In this system, the length of cyclic prefix is determined by the maximum physical time delay, and is the same as that in a normal OFDM system. Thus it is readily applicable to the environment in which the DOA is statistically independent of the time delay.

[0080] When the DOA of a path is statistically related to the path delay, e.g., in TR and HR environments, one can not only achieve beamforming gain and diversity gain simultaneously, but also reduce the cyclic prefix, thus obtaining improved spectrum efficiency.

[0081] A further example of the present invention utilises combined beamforming and transmit diversity for frequency selective fading channels for two ray (TR) models.

[0082] Suppose the physical channel follows a TR model. With the base station antenna array, the complex baseband representation of a wireless channel impulse response can be described as the following vector form

$$h_d(t;\tau) = \sum_{m=1}^{2} h_{d,m}(t)\delta(\tau - \tau_m) \tag{27}$$

with

$$h_{d,m}(t) = \sum_l \gamma_{m,l}(t) a_d(\theta_{m,l}) \tag{28}$$

where τ_m is the delay of the mth path resolved in time, $\gamma_{m,l}(t)$ and $a_d(\theta_{m,l})$ are the complex amplitude and downlink steering vector corresponding to lth DOA of the mth delay path. Because of the motion of the vehicular, $\gamma_{m,l}(t)$'s are wide-sense stationary (WSS) narrow band complex Gaussian processes, which are zero-mean and statistically independent for different m's, or l's. Suppose all $\gamma_{m,l}(t)$'s have the same normalized correlation function, $r(t)$ ($r(0) = 1$), but possibly different average power, $\sigma_{m,l}^2$, then

$$E\left[\gamma_{m,l}(t+\Delta t)\gamma_{m,l}^*(t)\right] = \sigma_{m,l}^2 r(\Delta t) \tag{29}$$

[0083] ISI exists when $\Delta\tau = \tau_2 - \tau_1$ is greater than the symbol interval. With combined beamforming and diversity technique, if the two rays are spatially separated, it is possible to transfer a frequency selective fading channel into a flat fading channel, yet maintain the transmit diversity.

[0084] Figure 6 shows a communication system with combined beamforming and transmit diversity for two-ray frequency selective fading channels. The signal to be transmitted, $s(n)$, is first coded in a coding module 3 using space-time codes, with the two branch outputs as $s_1(n)$ and $s_2(n)$. $s_1(n)$ is then fed through a delay 14 to delay $s(n)$ by $\Delta\tau$, yielding $x_1(n)$, which is further passed to transmit beamformer 11, (w_1). The second branch output $s_2(n)$ is directly passed to the other transmit beamformer 12, (w_2). The beamforming outputs are then combined in combiner 13 and sent by transmit antennas 1A, 1B, 2, yielding the transmitted signal as follows:

$$x(n) = w_1^H x_1(n) + w_2^H s_2(n) \quad (30)$$

[0085] The received signal, *y(n)*, at the mobile terminal single antenna 4 is given by

$$\begin{aligned} y(n) &= w_1^H h_{d,1} x_1(n) + w_1^H h_{d,2} x_1(n - \Delta\tau) \\ &+ w_2^H h_{d,1} s_2(n) + w_2^H h_{d,2} s_2(n - \Delta\tau) + w(n) \end{aligned} \quad (31)$$

[0086] Denoting $z(n) = y(n+\Delta\tau)$, and considering the pre-alignment of the two transmitted signals, gives:

$$\begin{aligned} z(n) &= w_1^H h_{d,1} s_1(n) + w_1^H h_{d,2} s_1(n - \Delta\tau) \\ &+ w_2^H h_{d,1} s_2(n + \Delta\tau) + w_2^H h_{d,2} s_2(n) + w(n + \Delta\tau) \end{aligned} \quad (32)$$

[0087] The beamforming weights are chosen such that the first branch output, $s_1(n)$, just goes through the first path, $h_{d,1}$between the base station antenna array and the receive antenna 4; while the second branch output, $s_2(n)$, just goes through the second path, $h_{d,2}$ between the base station antenna array and the receive antenna 4. Mathematically,

$$\begin{cases} w_1^H h_{d,2} = 0 \\ \left| w_1^H h_{d,1} \right|^2 = \max \end{cases}$$

and

$$\begin{cases} w_2^H h_{d,1} = 0 \\ \left| w_2^H h_{d,2} \right|^2 = \max \end{cases}$$

[0088] In this case the ISI terms are suppressed completely, and *z(n)* can be written as

$$z(n) = w_1^H h_{d,1} s_1(n) + w_2^H h_{d,2} s_2(n) + w(n + \Delta\tau) \quad (33)$$

[0089] Thus the frequency selective fading channel is now transformed into a flat fading channel, with which the transmit diversity method can be applied.

[0090] Conveniently, the transmit beamforming weights can be chosen by maximizing the average transmit SINR functions:

$$J_1(w_1) = \frac{w_1^H R_{d,1} w_1}{w_1^H R_{d,2} w_1} \text{ and } J_2(w_1) = \frac{w_2^H R_{d,2} w_2}{w_2^H R_{d,1} w_2}$$

where

$$R_{d,m} = E\left[h_{d,m}(t)h_{d,m}^H(t)\right] = \sum_l \sigma_{m,l}^2 a_d(\theta_{m,l}) a_d^H(\theta_{m,l}) \tag{34}$$

is the downlink channel covariance matrix of the mth path.

[0091] Preferably, the transmit beamforming weights can be chosen by maximizing the average receive SINR at the mobile receiver, i.e.,

$$J = \frac{w_1^H R_{d,1} w_1 + w_2^H R_{d,2} w_2}{w_1^H R_{d,2} w_1 + w_2^H R_{d,1} w_2 + \sigma_n^2}, \tag{35}$$

[0092] Advantageously, the transmit beamforming weights, w_m, can be chosen as the principal eigenvector of $R_{d,m}$.

[0093] Again, the frequency calibration method disclosed in Y-C. Liang and F. Chin, "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4 is used to estimate the DCCM from UCCM directly.

[0094] The above method for achieving combined beamforming and transmit diversity gain is called pre-alignment (PAL) method. The purpose of delaying $s_1(n)$ by $\Delta\tau$ is to make sure that the major components of the two sequences, $s_1(n)$ and $s_2(n)$ arrive at the receiver at the same time. Therefore, the delay spread has been reduced to zero. On the other hand, beamforming is used to minimize the ISI effect as well as to artificially generate two uncorrelated channels, with which the transmit diversity gain is achieved.

[0095] The PAL method requires the delay information, $\Delta\tau$, which is embedded in the downlink power-delay-DOA (PDD) profile. Even though the PDD profile is time varying, it changes slowly in time. Also, downlink PDD profile is almost the same as uplink PDD profile, which can be estimated from received uplink signals.

[0096] The PAL method can also be applied to the systems whose number of rays is grater than 2. In this case, it requires more than 2 branches of space-time coding outputs, and each output except the first one corresponds to one delay.

[0097] If the number of space-time coding outputs is fixed, say 2, the two major rays can be selected in order to generate the delay, $\Delta\tau$, and the transmit beamforming weights. The direct application of this system is to reduce inter-fmger-interference in CDMA as the total number of fingers is reduced.

[0098] Conventionally, when the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the delayed signal only goes through one ray $h_1(k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal only goes through another ray $h_2(k)$ between the base station multiple transmit antennae and the receive antenna.

[0099] Advantageously, when the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the average transmit SINR function at the base station is maximized for each ray.

[0100] Preferably, when the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the average receive SINR function at the mobile terminal is maximized.

[0101] Another example of the present invention utlises OFDM with combined beamforming and transmit diversity for frequency selective fading channels for two ray (TR) models.

[0102] There is a direct use of delay spread reduction in OFDM. In a typical OFDM system, a cyclic prefix is added in order to remove the ISI and to guarantee the orthogonality between each sub-channel. The length of the cyclic prefix should be greater than the maximum time delay, which can be as large as 40 μs for a mobile wireless communication environment. The adding of the cyclic prefix not only degrades the spectrum efficiency, but also occupies one portion of the transmit power. The spectrum efficiency and power efficiency of the OFDM system can be greatly improved if the cyclic prefix can be reduced while maintaining the same performance.

[0103] Suppose the physical channel follows a TR model with parameters $(\alpha_k,\theta_k,\tau_k)$, k = 1,2 and $\tau_1<\tau_2$. α_k's are statistically independent, zero mean complex Gaussian processes with variance σ_k^2. ISI exists when $\Delta\tau = \tau_2-\tau_1$ is greater than the inverse of bandwidth.

[0104] Figure 7 illustrates an OFDM system with combined beamforming and transmit diversity for TR models embodying the present invention. The transmitted signal at the kth tone of the nth block, $S(n;k)$, is first coded using space-time codes in coding module 3, yielding two branch outputs, $S_1(n;k)$ and $S_2(n;k)$. Both branch outputs $S_1(n;k)$ and $S_2(n;k)$ are passed into respective OFDM transmit processors 8,9 without adding cyclic prefixes. $S_1(n;k)$ is then delayed

in delay 14 by $\Delta\tau$, yielding $X_1(n;k)$, which is further passed to transmit beamformer 11, (w_1). The second branch output $S_2(n;k)$ is directly passed to the other transmit beamformer 12, (w_2). The beamforming outputs are then combined and sent on the base station transmit antenna array 1A, 1B, 2, yielding the transmitted signal as follows:

$$x(n;k) = w_1^H x_1(n;k) + w_2^H s_2(n;k) \tag{36}$$

[0105] At the mobile terminal single antenna 4, the received signals are first passed into a normal OFDM receive processor 10. The beamforming weights are chosen such that the first branch output, $S_1(n;k)$ or its inverse FFT (IFFT), $s_1(n;k)$, just goes through the first path, $h_1(n;k)$ between the base station antenna array and the receive antenna 4; while the second branch output, $S_2(n;k)$ or its inverse FFT (IFFT), $s_2(n;k)$, just goes through the second path, $h_1(n;k)$ between the base station antenna array and the receive antenna 4. Once the transmit beamforming weights are properly chosen, the FFT output of the received signal at the mobile station becomes

$$Z[n;k] = w_1^H H_1[n;k]S_1[n;k] + w_2^H H_2[n;k]S_2[n;k] + W[n;k + \lfloor \Delta\tau f_s \rfloor] \tag{37}$$

[0106] Comparing equation (37) with equation (5), with the aid of downlink beamforming, two different channels have been artificially created which can be space-time decoded by module 5 to recover the transmitted signal. Further, permutation decoding method can be easily applied if $S_1(n;k)$ and $S_2(n;k)$ are chosen as follows.

	$t=k$	$t=k+1$
$S_1(n;t)$	$s(n;k)/\sqrt{2}$	$s^*(n;k+1)/\sqrt{2}$
$S_2(n;t)$	$s(n;k+1)/\sqrt{2}$	$-s^*(n;k)/\sqrt{2}$

[0107] When PAL is applied to an OFDM system with combined beamforming and transmit diversity for TR models, it is not necessary to add the cyclic prefix. Thus benefiting from the advantages of: transmit diversity; beamforming gain; and increased spectrum efficiency.

[0108] Conveniently, the transmit beamforming weights can be chosen by maximizing the average transmit SINR functions.

[0109] Preferably, the transmit beamforming weights can be chosen by maximizing the average receive SINR at the mobile receiver.

[0110] Advantageously, the transmit beamforming weights, w_m, can be chosen as the principal eigenvector of $R_{d,m}$.

[0111] Again, the frequency calibration method disclosed in Y-C. Liang and F. Chin, "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4 is used to estimate the DCCM from UCCM directly.

[0112] A comparison of the spectrum efficiency and power savings by using this delay spread reduction method will follow.

[0113] A further example of the invention utilises OFDM with combined beamforming and transmit diversity for frequency selective fading channels for hilly terrain (HT) models.

[0114] Even though the maximum time delay can be as large as 40 µs, a wireless channel satisfying HT model can be described by several dominated clustered paths, each of which has a small delay spread. These clustered paths are also spatially separated. For an OFDM with typical HT power-delay profile whose maximum time delay is 20 µs, and maximum delay spread for each clustered path is 2 µs, the minimum length of cyclic prefix is 20µs in order to remove the ISI. However, with the PAL method, the cyclic prefix duration can be reduced to 2 µs .

[0115] Suppose the two clustered paths are delayed by ψ,and for simplicity, assume the delay spread for each clustered path is $\Delta\psi$. The impulse response of the time varying channel can be described as

$$h(t;\tau) = h_1(t;\tau)\left[u(\tau) - u(\tau - \Delta\psi)\right] + h_2(t;\tau - \psi)\left[u(\tau - \psi) - u(\tau - \psi - \Delta\psi)\right] \quad (38)$$

where $h_1(t;\tau)$ and $h_2(t;\tau)$ correspond to the channel responses of the first and second clustered paths, respectively; and $u(x)$ is a unit step function..

[0116] Figure 8 shows an OFDM system embodying the present invention with combined beamforming and transmit diversity for hilly terrain (HT) model in encoder module 3. The signal to be transmitted at the kth tone of the nth block, $S(n;k)$, is first coded using space-time codes in encoder module 3, yielding two branch outputs, $S_1(n;k)$ and $S_2(n;k)$ which are passed into respective normal OFDM transmit processors 8,9, whose cyclic prefix length is $\Delta\psi$, instead of $\psi + \Delta\psi$. The output from the first branch is then delayed by ψ in delay 15, while the output from the second branch remains unchanged. After that, the signals are passed into respective transmit beamformers 11,12, (w_1 and w_2), respectively. The beamforming outputs are then combined in combiner 13, and transmitted out through the base station transmit antenna array 1A, 1B, 2.

[0117] The beamforming weights are chosen such that the first branch input just goes through the first clustered path, while the second branch input just goes through the second clustered path - i.e. the beamforming weights are chosen such that the first branch output, $s_1(n)$, just goes through the first path, $h_{d,1}$ between the base station antenna array and the receive antenna 4; while the second branch output, $s_2(n)$, just goes through the second path, $h_{d,2}$ between the base station antenna array and the receive antenna 4. The signals received at the mobile terminal single antenna 4 are first passed into a normal OFDM receive processor 10, followed by a space-time decoding module 5. Within the normal OFDM receive processor 10, the received signal after FFT becomes

$$Z[n;k] = w_1^H H_1[n;k]S_1[n;k] + w_2^H H_2[n;k]S_2[n;k] + W\left[n;k + \lfloor \psi f_r \rfloor\right] \quad (39)$$

where $\lfloor x \rfloor$ denotes the maximum integer which is not greater than x. Comparing equation (39) with equation (5), with the aid of downlink beamforming, two different channels have been artificially generated, which are space-time decoded to recover the transmitted signal. Permutation decoding methods can be easily applied if $S_1(n;k)$ and $S_2(n;k)$ are chosen as follows.

	$t=k$	$t=k+1$
$S_1(n;t)$	$s(n;k)/\sqrt{2}$	$s^*(n;k+1)/\sqrt{2}$
$S_2(n;t)$	$s(n;k+1)/\sqrt{2}$	$-s^*(n;k)/\sqrt{2}$

[0118] Conveniently, the transmit beamforming weights can be chosen by maximizing the average transmit SINR functions.

[0119] Preferably, the transmit beamforming weights can be chosen by maximizing the average receive SINR at the mobile receiver.

[0120] Advantageously, the transmit beamforming weights, w_m, can be chosen as the principal eigenvector of $R_{d,m}$.

[0121] As previously mentioned, there follows a comparison the spectrum efficiency of a OFDM system with different cyclic prefix lengths.

[0122] The parameters are Bandwidth B = 800 kHz, maximum time delay = 40. For HT models, the maximum delay spread for each clustered path is 5. To make the tones orthogonal to each other, the symbol duration is N/B, where N is the number of tones in each OFDM symbol. The total block length is the summation of the symbol duration and the additional guard interval, which is 40, 5, and 0 for OFDM without PAL, HT with PAL and TR with PAL, respectively.

[0123] Table I illustrates the uncoded transmit data rate for OFDM systems with different number of tones using QPSK modulation. It is seen that, for a given modulation scheme and with the same number of tones, the transmit data rate can increase to 1.6Mbps for TR environments by using PAL, independent of the N value. For HT with PAL, the spectrum efficiency is also increased as compared with that without PAL.

Table I:

transmit data rate comparison			
	N=128	N=64	N=32
Without PAL	1.28 Mbps	1.07 Mbps	800 kbps
HT with PAL	1.55 Mbps	1.51 Mbps	1.42 Mbps
TR with PAL	1.6 Mbps	1.6 Mbps	1.6 Mbps

[0124] Here follows a comparison of the power savings for OFDM with different lengths of cyclic prefix:

[0125] Due to the adding of a cyclic prefix, the effective $\frac{E_b}{N_0}$ is smaller than the actual transmit $\frac{E_b}{N_0}$. With delay spread reduction, the transmit power is more efficiently used. Table II illustrates the power savings for OFDM systems with delay spread reduction using PAL for different number of tones in each OFDM block, as compared to normal OFDM systems.

Table II:

Power savings			
	N=128	N=64	N=32
HT with PAL	0.84 dB	1.5 dB	2.5 dB
TR with PAL	0.97 dB	1.76 dB	3.0 dB

Beamforming and diversity gain:

[0126] With combined beamforming and diversity gain, it takes less $\frac{E_b}{N_0}$ in order for the system to achieve a given bit-error-rate (BER) requirement. Alternatively, the beamforming and diversity gain can be translated to larger spectrum efficiency using higher modulation scheme such as 128 QAM or 256 QAM.

[0127] A further embodiment of the present invention relates to adaptive delay spread reduction with combined beamforming and diversity gain:

[0128] The previously described embodiments are designed for different environments. In real applications, the power-delay-DOA (PDD) profile may change with respect to time due to the motion of a vehicle, thus the delay spread reduction scheme should follow this variation accordingly in order to achieve maximum spectrum efficiency. Figure 9 shows an OFDM system with combined beamforming, transmit diversity and adaptive delay spread reduction for downlink embodying the present invention. The OFDM system of Figure 9 comprises the system of Figure 8 but supplemented by UCCM estimation and power-delay-DOA profile estimation. Thus, in addition to the functionality provided by the system of Figure 8, this system has the following functionality.

- From uplink signals received at the base station, the time-delay and direction-of-arrival (DOA) information is estimated for each received path, using training sequences or blind techniques. Based on the estimated time-delay and DOA information, uplink power-delay-DOA (PDD) profile, and each clustered path's UCCM are estimated;
- Based on uplink PDD profile, the following parameters are determined: diversity order, time delays for each clustered path, and the maximum delay spread for the clustered paths.
- The uplink PDD profile is used to design the adaptive delay reduction scheme, thus the adaptive cyclic prefix adding scheme;
- Each clustered path's DCCM is estimated from its corresponding UCCM using FC method disclosed in Y-C. Liang and F. Chin "Downlink beamforming methods for capacity enhancement in wireless communication systems", Singapore Patent Application No. 9904733.4, then applied, together with time delay information, for constructing transmit beamforming weights;
- The base station informs the MS the length of added cyclic prefix;
- Adaptive modulation is also used to further improve the spectrum efficiency based on the diversity order/channel condition. Specifically, based on uplink PDD profile, the maximum achievable diversity order is determined. If the achievable diversity order is large, a higher modulation scheme is applied; otherwise, a smaller modulation scheme is applied.

[0129] It should be noted that the number of branch outputs after space-time coding in module 3 can be greater than two, depending on the diversity order to be achieved.

[0130] The above description considers the combined beamforming, transmit diversity and delay spread reduction implemented at the base station. In fact, multiple diversity antennas can be added at the mobile terminal as well to achieve receive diversity. In this case, larger diversity gains can be achieved:

[0131] Even though OFDM is used to show how the delay spread can be reduced, while yet maintaining beamforming and transmit diversity gain, the disclosure in this application can be applied to other multi-carrier modulation schemes, such as MC-CDMA, MC-DS-CDMA and single carrier systems with cyclic prefix.

[0132] In a multiuser environment, the beamforming weights can be generated by considering all users' channel/ DOA information; therefore, the disclosure in this application is applicable in different multiple access schemes, such as time-division-multiple-access (TDMA), frequency-division-multiple-access (FDMA), and code-division-multiple-access (CDMA).

"comprising" means "including or consisting of".

[0133] The features disclosed in the foregoing description, or the following claims, or the accompanying drawings, expressed in their specific forms or in terms of a means for performing the disclosed function, or a method or process for attaining the disclosed result, as appropriate, may, separately, or in any combination of such features, be utilised for realising the invention in diverse forms thereof.

Claims

1. A method of achieving transmit diversity gain for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with at least a single receive antenna, the method comprising the steps of:

   providing a signal to be transmitted $s(n)$;
   space-time encoding the signal $s(n)$ to produce at least two separate signals $s_1(n),s_2(n)$, each on a respective output;
   feeding each output signal $s_1(n),s_2(n)$ to a zero-forcing pre-equaliser having a respective function $g_1(k)$, $g_2(k)$ to produce an output signal $x_1(n)$, $x_2(n)$;
   feeding the output signal $x_1(n)$, $x_2(n)$ of each pre-equaliser to a transmit antenna;
   transmitting the output signals $x_1(n)$, $x_2(n)$ over respective physical channels $h_1(k)$, $h_2(k)$;
   receiving the output signals $x_1(n)$, $x_2(n)$ at at least a single receive antenna; and space-time decoding the received signals, wherein
   the functions $g_1(k)$, $g_2(k)$ of the zero-forcing pre-equalisers are selected such that the channel responses $g_1(k)^*h_1(s)$, $g_2(k)^*h_2(k)$ of the respective physical channels $h_1(k)$, $h_2(k)$ are flat fading channels.

2. A method according to Claim 1, wherein the communications system is a time-division duplex system and the method includes the further step of deriving the real channel coefficients from uplink channel coefficients for use in selecting the functions $g_1(k)$, $g_2(k)$ of the pre-equalisers.

3. A method according to Claim 2, wherein the step of deriving the real channel coefficients from uplink channel coefficients uses training symbols from the uplink channel.

4. A method according to Claim 2, wherein the step of deriving the real channel coefficients from uplink channel coefficients uses blind techniques.

5. A method according to Claim 1, wherein the communications system is a frequency-division duplex system and the method includes the further step of deriving the real channel coefficients by sending a set of training symbols to the receive antenna of the mobile terminal, the mobile terminal estimating the real channel coefficients and feeding back channel coefficient information to the base station.

6. A base station with multiple transmit antennae for communicating with a mobile terminal having at least a single receive antenna over physical channels $h_1(k)$, $h_2(k)$ the base station comprising:

   a space-time encoder having an input of a signal to be transmitted $s(n)$ and at least two outputs each producing a separate signal $s_1(n),s_2(n)$;
   at least two zero-forcing pre-equalisers, each fed by a respective output signal $s_1(n),s_2(n)$ and having a respective function $g_1(k)$, $g_2(k)$ to produce an output signal $x_1(n)$, $x_2(n)$; and
   at least two transmit antennae, each being fed by the output signal $x_1(n)$, $x_2(n)$ of a respective one of the pre-

equalisers, wherein the functions $g_1(k)$, $g_2(k)$ of the zero-forcing pre-equalisers are selected such that the channel responses $g_1(k)^*h_1(k)$, $g_2(k)^*h_2(k)$ of the respective physical channels $h_1(k)$, $h_2(k)$ are flat fading channels.

7. A communications system comprising the base station of Claim 6 and a mobile terminal having at least a single receive antenna and a space-time decoder to decode the signals received from the base station.

8. A method of achieving combined beamforming and transmit diversity for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with at least a single receive antenna, the method comprising the steps of:

providing a signal to be transmitted $S(n;k)$;
space-time encoding the signal $S(n;k)$ to produce at least two separate signals $S_1(n;k)$,$S_2(n;k)$, each on a respective output;
feeding each output signal $S_1(n;k)$,$S_2(n;k)$ to a transmit processor to produce an output signal $X_1(n;k)$, $X_2(n;k)$;
applying respective selected transmit beamforming weights to each output signal $X_1(n;k)$, $X_2(n;k)$;
feeding the respective weighted signals to a signal combiner to perform a summing function of the signals and produce a signal X(n;k) for transmission;
feeding the summed signal X(n;k) to each of the multiple transmit antennae for transmission;
transmitting the signals X(n;k) over physical channel $h(n;k)$;
receiving the received signal Y(n;k) at at least a single receive antenna;
feeding the received signal Y(n;k) to a receive processor to produce an output signal; and
space-time decoding the received signal.

9. A method according to Claim 8, wherein the respective transmit beamforming weights are selected as the eigenvectors corresponding to the two largest eigenvalues of the downlink channel covariance matrix (DCCM) of the physical channels h(n;k).

10. A method according to Claim 8, wherein the physical channel h(n;k) consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, the transmit processors do not add cyclic prefixes and one of the output signals from the transmit processors is delayed by $\Delta\tau$ before the respective selected transmit beamforming weight is applied thereto.

11. A method according to Claim 8, wherein the physical channel h(n;k) consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, the beamforming weights being chosen such that the delayed signal or its inverse fast Fourier transform (IFFT) only goes through one channel $h_1(n;k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal or its IFFT only goes through another channel $h_2(n;k)$ between the base station multiple transmit antennae and the receive antenna, thereby creating two different channels which can be space-time decoded to recover the transmitted signal.

12. A method according to Claim 8, wherein the physical channel h(n;k) consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, the beamforming weights being chosen such that the average transmit SINR function at the base station is maximized for each ray.

13. A method according to Claim 8, wherein the physical channel h(n;k) consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, the beamforming weights being chosen such that the average receive SINR function at the mobile terminal is maximized.

14. A method according to Claim 8, wherein the physical channel h(n;k) consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, the beamforming weights for each ray are chosen as the principal eigenvector of the downlink channel covariance matrix (DCCM) corresponding to that ray.

15. A method according to Claim 8, wherein the physical channel h(n;k) consists of two time-delayed clustered rays, $h_1(n;k)$ and $h_2(n;k)$, with delay ψ, and maximum excess delay for the clusters $\Delta\psi$, the transmit processors have a cyclic prefix length of $\Delta\psi$ and one of the output signals from the transmit processors is delayed by ψ before the respective selected transmit beamforming weight is applied thereto.

16. A method according to Claim 15, wherein the beamforming weights are chosen such that the delayed signal or its inverse fast Fourier transform (IFFT) only goes through one channel $h_1(n;k)$ between the base station multiple

transmit antennae and the receive antenna, whilst the undelayed signal or its IFFT only goes through another channel $h_2(n;k)$ between the base station multiple transmit antennae and the receive antenna, thereby creating two different channels which can be space-time decoded to recover the transmitted signal.

17. A method according to Claim 15, wherein the beamforming weights being chosen such that the average transmit SINR function at the base station is maximized for each clustered ray.

18. A method according to Claim 15, wherein the beamforming weights being chosen such that the average receive SINR function at the mobile terminal is maximized.

19. A method according to Claim 15, wherein the beamforming weights for each clustered ray are chosen as the principal eigenvector of the downlink channel covariance matrix (DCCM) corresponding to that clustered ray.

20. A method according to Claim 15, comprising the further steps of:

estimating a power-delay-DOA profile for channel $h(n;k)$; and, based on the profile: determining the cyclic prefix, $\Delta\psi$, to be added by the transmit processors; determining the delay ψ; diversity order and modulation scheme; and determining the transmit beamforming weights.

21. A method according to Claim 20, comprising the further step of estimating the downlink channel covariance matrix (DCCM) from the uplink channel covariance matrix (UCCM) to construct transmit beamforming weights.

22. A method according to Claim 21, comprising the further step of determining the diversity order and modulation scheme based on the profile.

23. A method according to Claim 8, wherein the transmit and receive processors are selected from the group consisting of: OFDM, MC-CDMA MC-DS-CDMA and a single carrier system with cyclic prefix.

24. A base station with multiple transmit antennae for communicating with a mobile terminal having at least a single receive antenna over physical channel $h(k)$, the base station comprising:

a space-time encoder having an input of a signal to be transmitted and at least two outputs each producing a separate signal;
at least two transmit processors each receiving one of the outputs from a respective space-time encoder;
at least two transmit beamformers each receiving an output from a respective transmit processor and applying a transmit beamforming weight thereto;
a signal combiner receiving signals from the beamformers and operable to perform a summing function of the signals from the beamformers and produce a signal for transmission by the multiple transmit antennae.

25. A base station according to Claim 24, wherein the physical channel $h(n;k)$ consists of two time-delayed rays, $h_1(n;k)$ and $h_2(n;k)$, with delay $\Delta\tau$, further comprising a delay of $\Delta\tau$ interposed between one of the multiple access transmit processor outputs and a beamformer to delay the signal output from the transmit processor by $\Delta\tau$ before the respective selected transmit beamforming weight is applied thereto, wherein the transmit processors do not add cyclic prefixes.

26. A base station according to Claim 24, wherein the physical channel $h(n;k)$ consists of two time-delayed clustered rays, $h_1(n;k)$ and $h_2(n;k)$, with delay ψ and maximum excess delay for the clusters $\Delta\psi$, further comprising a delay of ψ interposed between one of the multiple access transmit processor outputs and a beamformer to delay the signal output from the transmit processor by ψ before the respective selected transmit beamforming weight is applied thereto, the transmit processors having a cyclic prefix length of $\Delta\psi$.

27. A base station according to Claim 24, further comprising a first processor to determine a power-delay-DOA profile estimate for channel $h(n;k)$; and, based on the profile, determine: the length, $\Delta\psi$, of the cyclic prefix to be added by the transmit processors; the delay ψ; diversity order and modulation scheme; and the transmit beamforming weights.

28. A base station according to Claim 27, further comprising a second processor to estimate a downlink channel covariance matrix (DCCM) from the uplink channel covariance matrix (UCCM) to construct transmit beamforming

weights.

29. A base station according to Claim 15, wherein the transmit and receive processors are selected from the group consisting of: OFDM, MC-CDMA MC-DS-CDMA and single carrier system with cyclic prefix.

30. A communications system comprising the base station of Claim 24 and a mobile terminal having at least a single receive antenna, a receive processor to produce an output signal and a space-time decoder to decode the output signal.

31. A method of achieving combined beamforming and transmit diversity for frequency selective fading channels in a communication system having a base station with multiple transmit antennae and a mobile terminal with at least a single receive antenna, the method comprising the steps of:

providing a signal to be transmitted $s(n)$;
space-time encoding a signal to be transmitted $s(n)$ to produce at least two separate signals $s_1(n), s_2(n)$, each on a respective output;
delaying one of the space-time encoded output signals by $\Delta\tau$;
applying respective selected transmit beamforming weights to the delayed and undelayed signals;
feeding the respective weighted signals to a signal combiner to perform a summing function of the signals and produce a signal for transmission;
feeding the summed signal to each of the multiple transmit antennae for transmission;
transmitting the summed signals over the physical channel h(k);
receiving the major components of the transmitted signals at at least a single receive antenna at substantially the same time; and
space-time decoding the received signal.

32. A method according to Claim 31, wherein the physical channel h(k) consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the beamforming weights are chosen such that the delayed signal only goes through one ray $h_1(k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal only goes through another ray $h_2(k)$ between the base station multiple transmit antennae and the receive antenna.

33. A method according to Claim 31, wherein the physical channel h(k) consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the beamforming weights are chosen such that the average transmit SINR function at the base station is maximized for each ray.

34. A method according to Claim 31, wherein the physical channel h(k) consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the beamforming weights are chosen such that the average receive SINR function at the mobile terminal is maximized.

35. A method according to Claim 31, wherein the physical channel h(k) consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the beamforming weights for each ray are chosen as the principal eigenvector of the downlink channel covariance matrix (DCCM) corresponding to that ray.

36. A method according to Claim 31, wherein the physical channel h(k) consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the delay $\Delta\tau$ is derived from downlink channel information.

37. A method according to Claim 31, wherein the physical channel h(k) consists of two time-delayed rays $h_1(k)$, $h_2(k)$ with delay $\Delta\tau$, the delay $\Delta\tau$ is derived from uplink channel information.

38. A method according to Claim 31, wherein the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the delayed signal only goes through one ray $h_1(k)$ between the base station multiple transmit antennae and the receive antenna, whilst the undelayed signal only goes through another ray $h_2(k)$ between the base station multiple transmit antennae and the receive antenna.

39. A method according to Claim 31, wherein the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the average transmit SINR function at the base station is maximized for each ray.

40. A method according to Claim 31, wherein the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights are chosen such that the average receive SINR function at the mobile terminal is maximized.

41. A method according to Claim 31, wherein the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the beamforming weights for each ray are chosen as the principal eigenvector of the downlink channel covariance matrix (DCCM) corresponding to that ray.

42. A method according to Claim 31, wherein the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the delay $\Delta\tau$ is derived from downlink channel information.

43. A method according to Claim 31, wherein the physical channel h(k) consists of multiple rays with two major rays $h_1(k)$, $h_2(k)$ delayed by $\Delta\tau$, the delay $\Delta\tau$ is derived from uplink channel information.

44. A base station with multiple transmit antennae for communicating with a mobile terminal having at least a single receive antenna over physical channel $h(k)$ having two time-delayed rays, $h_1(k)$ and $h_2(k)$, the base station comprising:

a space-time encoder having an input of a signal to be transmitted and at least two outputs each producing a separate signal;
at least two transmit beamformers each receiving an output from the space-time encoder and applying a transmit beamforming weight thereto;
a signal combiner receiving signals from the beamformers and operable to perform a summing function of the signals from the beamformers and produce a signal for transmission by each of the multiple transmit antennae, wherein a delay of $\Delta\tau$ is interposed between the space-time encoder and one of the beamformers such that the major components of the transmitted signals are received at at least a single receive antenna at substantially the same time.

45. A communications system comprising the base station of Claim 24 and a mobile terminal having at least a single receive antenna and a space-time decoder to decode the received signal.

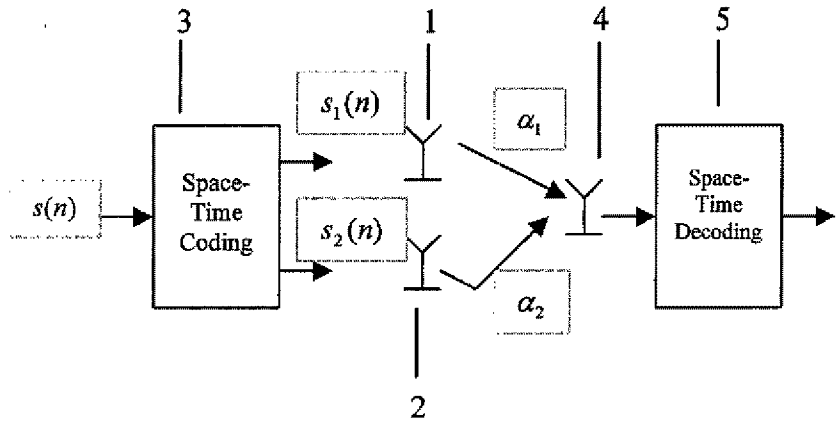


Figure1: Prior Art

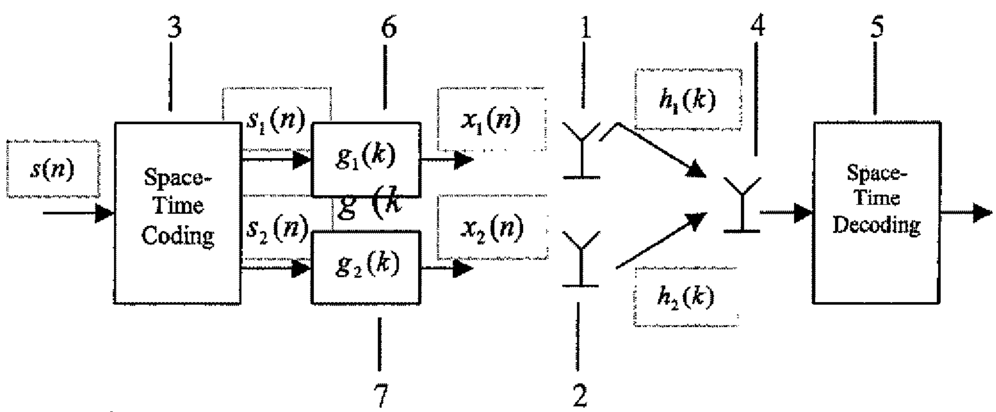


Figure2

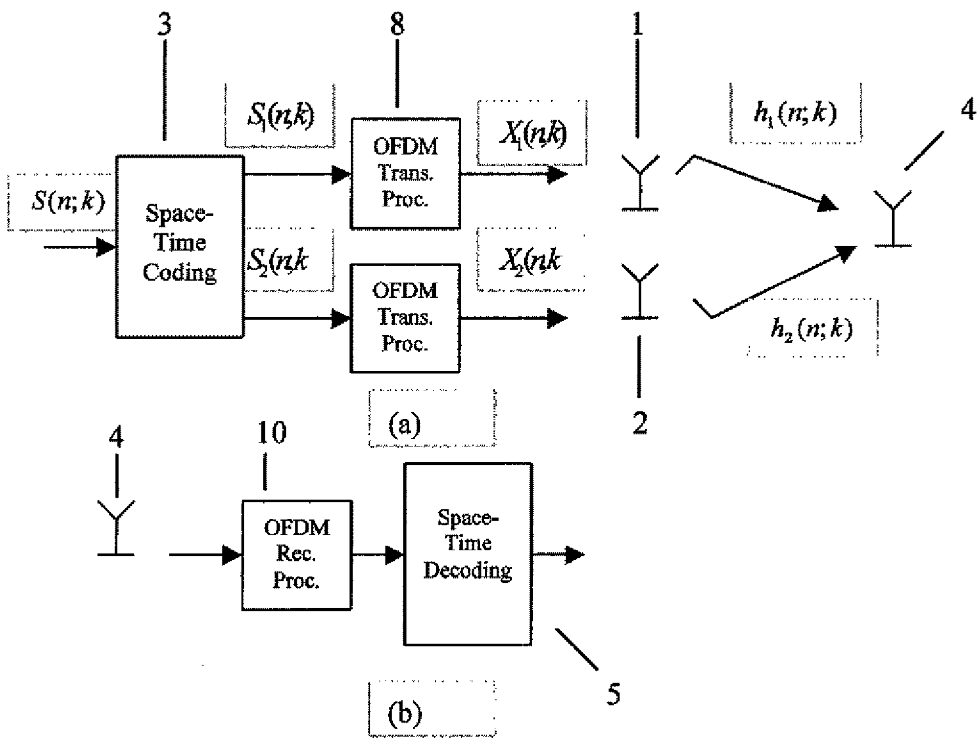


Figure3: Prior Art

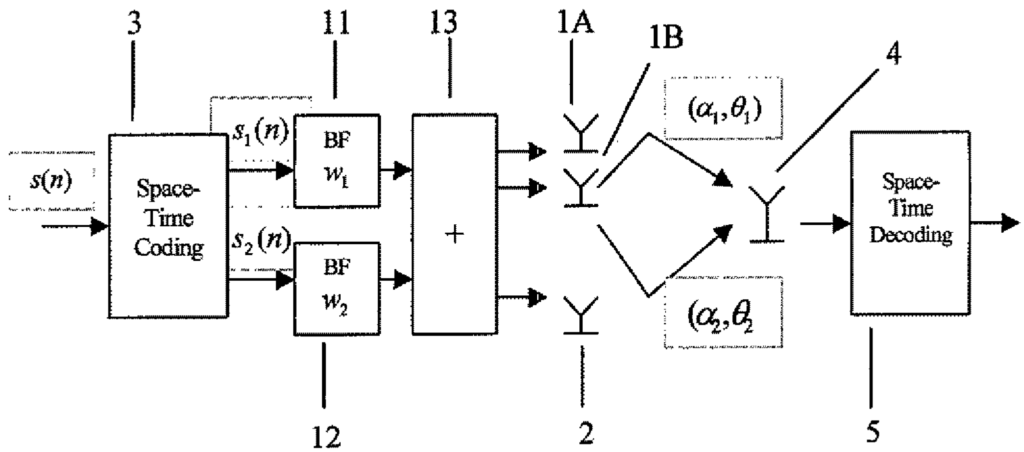


Figure4: Prior Art

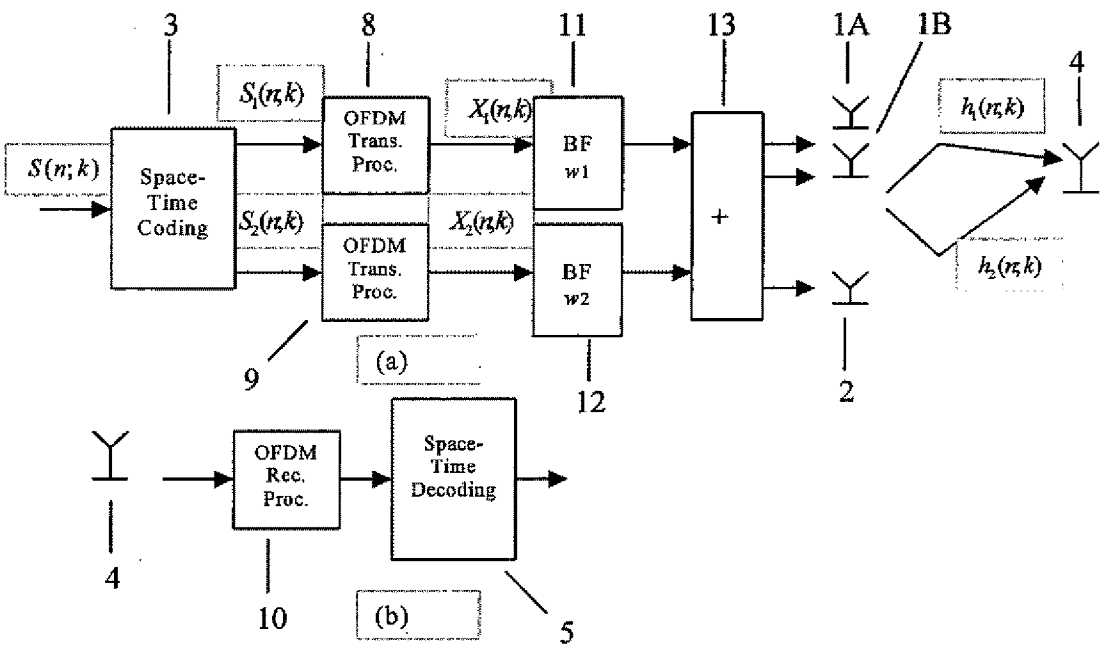

3
8
11
13
1A
1B
4
$S_1(n,k)$
$X_1(n,k)$
$h_1(n,k)$
$S(n;k)$
Space-Time Coding
OFDM Trans. Proc.
BF w1
+
$S_2(n,k)$
$X_2(n,k)$
OFDM Trans. Proc.
BF w2
$h_2(n,k)$
9
(a)
12
2
OFDM Rec. Proc.
Space-Time Decoding
4
10
(b)
5

Figure5

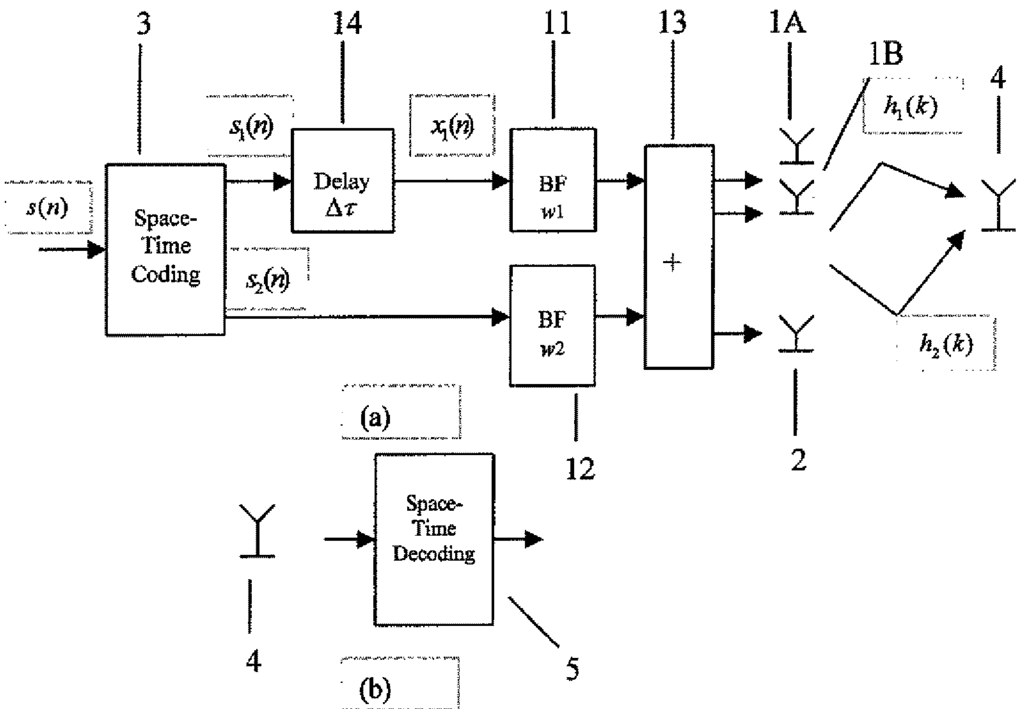

3
14
11
13
1A
1B
4
$s(n)$
$s_1(n)$
$x_1(n)$
$h_1(k)$
Space-Time Coding
Delay $\Delta\tau$
BF $w1$
BF $w2$
+
$s_2(n)$
$h_2(k)$
(a)
12
2
Space-Time Decoding
4
5
(b)

Figure6

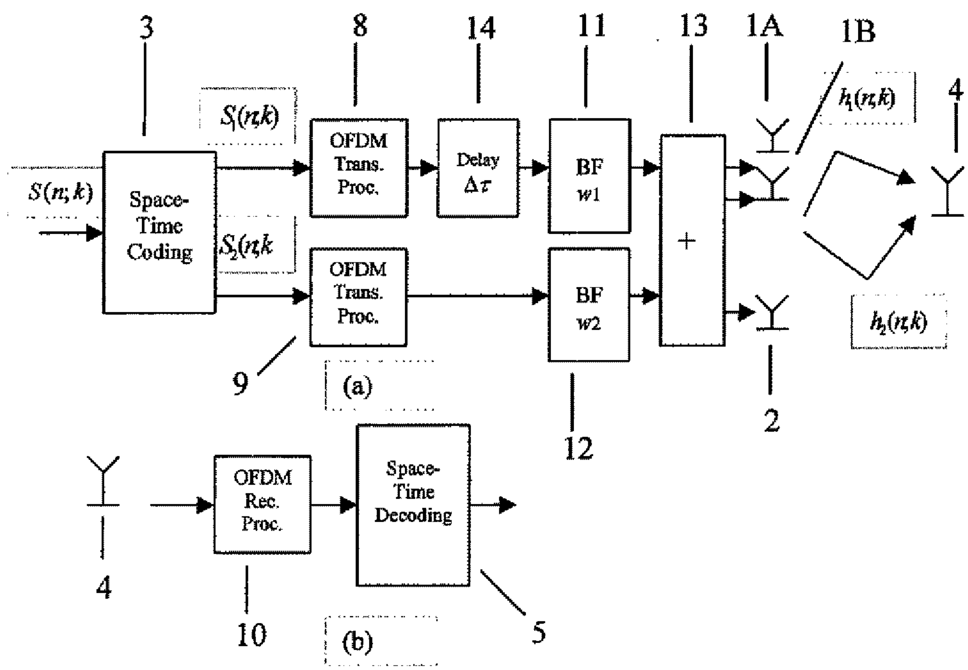

3
8
14
11
13
1A
1B
$S_1(n,k)$
$h_1(n,k)$
4
$S(n;k)$
Space-Time Coding
OFDM Trans. Proc.
Delay $\Delta\tau$
BF $w1$
+
$S_2(n,k$
OFDM Trans. Proc.
BF $w2$
$h_2(n,k)$
9
(a)
12
2
OFDM Rec. Proc.
Space-Time Decoding
4
10
(b)
5

Figure7

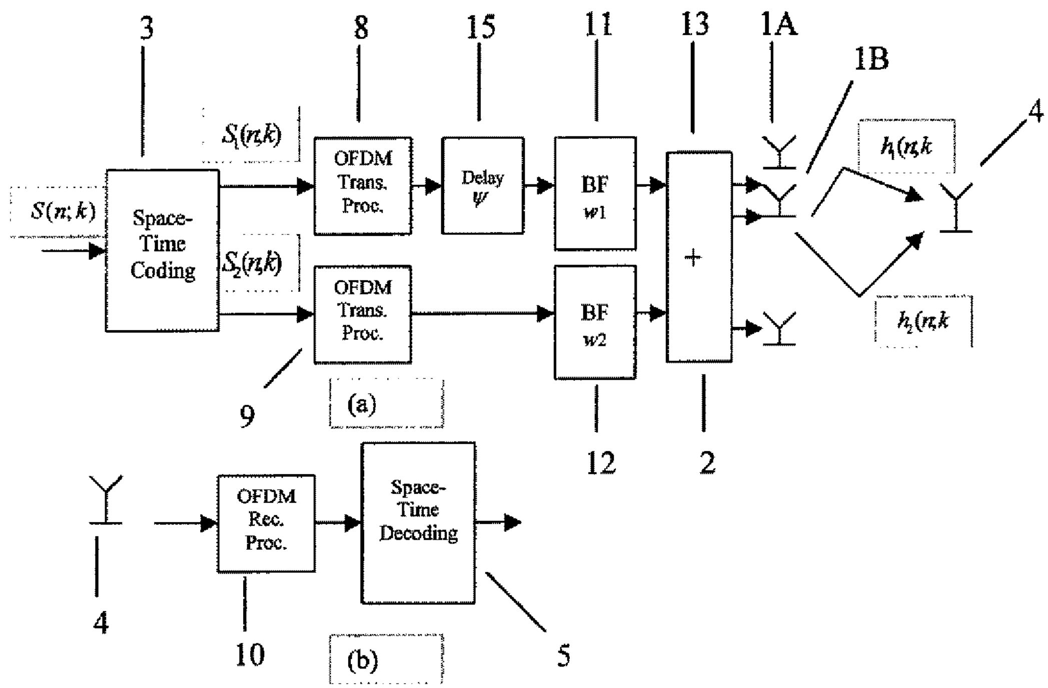
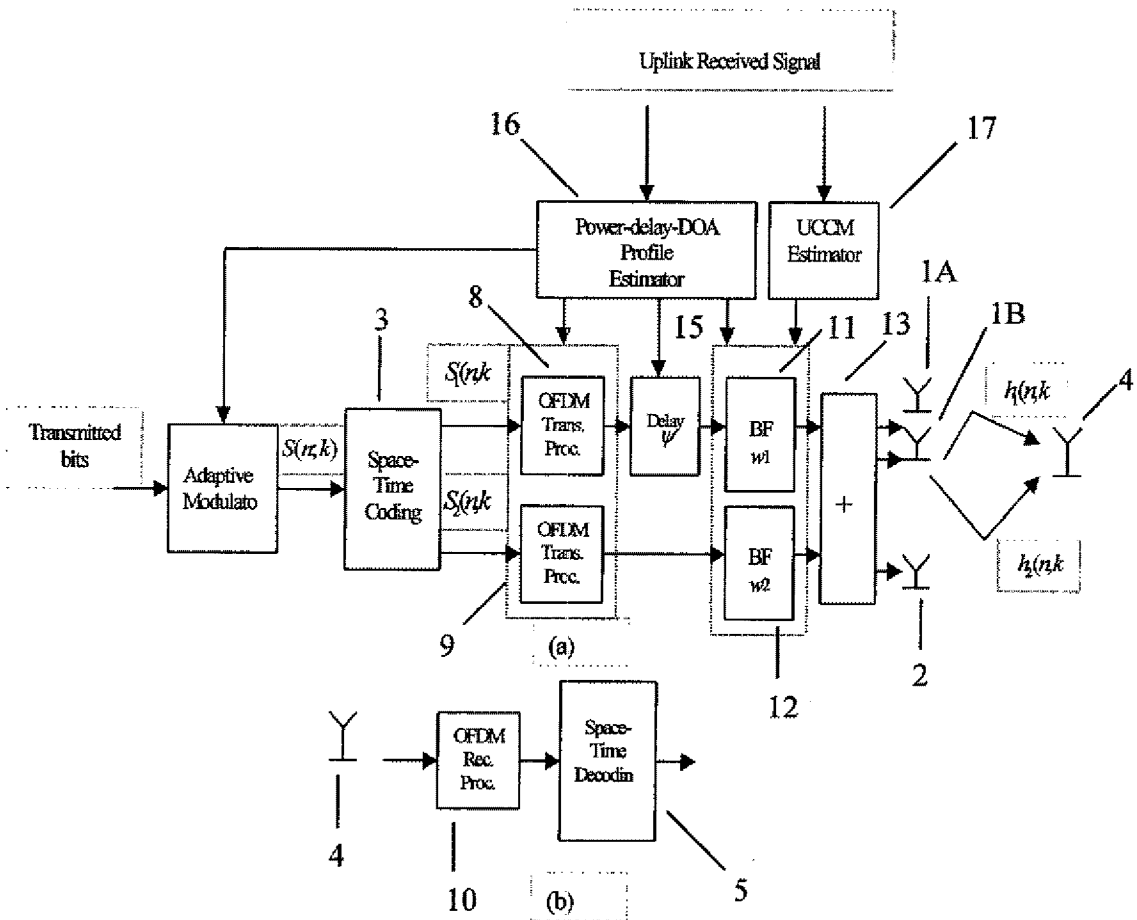

3
8
15
11
13
1A
1B
4
S1(n,k)
S(n;k)
Space-Time Coding
S2(n,k)
OFDM Trans. Proc.
Delay ψ
BF w1
+
h1(n,k
OFDM Trans. Proc.
BF w2
h2(n,k
9
(a)
12
2
OFDM Rec. Proc.
Space-Time Decoding
4
10
(b)
5

Figure8



Uplink Received Signal
16
17
Power-delay-DOA Profile Estimator
UCCM Estimator
1A
1B
8
3
15
11
13
4
Transmitted bits
Adaptive Modulato
S(n,k)
Space-Time Coding
OFDM Trans. Proc.
Delay
BF w1
BF w2
+
OFDM Trans. Proc.
9
(a)
2
12
OFDM Rec. Proc.
Space-Time Decodin
4
10
(b)
5

Figure9

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 02 25 4685

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
X	SAMPATH H ET AL: "Pre-equalization for MIMO wireless channels with delay spread" VTC 2000-FALL; VEHICULAR TECHNOLOGY CONFERENCE. BOSTON, MA, US, vol. 3 OF 6. CONF. 52, 24 - 28 September 2000, pages 1175-1178, XP010524685 IEEE ISBN: 0-7803-6508-9 * section 1 * * section 2.1 * * sections 3.1, 3.2, 3.3 * ---	1-7	H04L1/06 H04L25/03
X	LIU Z ET AL: "Transmit-antennae space-time block coding for generalized OFDM in the presence on unknown multipath" IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, IEEE INC. NEW YORK, US, vol. 19, no. 7, July 2001 (2001-07), pages 1352-1364, XP001112325 ISSN: 0733-8716 * abstract * * section II * * figure 1 * ---	1-7	TECHNICAL FIELDS SEARCHED (Int.Cl.7) H04L
A	YANG H ET AL: "An adaptive channel precoded space-time transmitter for 3GPP TDD system" GLOBECOM '01. 2001 IEEE GLOBAL TELECOMMUNICATIONS CONFERENCE. SAN ANTONIO, TX, IEEE GLOBAL TELECOMMUNICATIONS CONFERENCE, NEW YORK, NY: IEEE, US, vol. 1 OF 6, 25 - 29 November 2001, pages 529-532, XP001090311 ISBN: 0-7803-7206-9 * section II-B * * figure 3 * --- -/--	1-7	

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	10 April 2003	Ghigliotti, L

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
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A : technological background
O : non-written disclosure
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T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 02 25 4685

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
A,D	ALAMOUTI S M: "A simple transmit diversity technique for wireless communications" IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, IEEE INC. NEW YORK, US, vol. 16, no. 8, October 1998 (1998-10), pages 1451-1458, XP002100058 ISSN: 0733-8716 * the whole document *	1,6	
A,D	TAROKH V ET AL: "Space-time codes for high data rate wireless communication: performance criterion and code construction" IEEE TRANSACTIONS ON INFORMATION THEORY, IEEE INC. NEW YORK, US, vol. 44, no. 2, 1 March 1998 (1998-03-01), pages 744-765, XP002089112 ISSN: 0018-9448 * the whole document *	1-7	
			TECHNICAL FIELDS SEARCHED (Int.Cl.7)
X	WO 02 01732 A (NOKIA CORP ;NOKIA INC (US)) 3 January 2002 (2002-01-03) * page 42, line 7 - line 19 * * figures 11,26 *	8-45	
X	WO 01 78254 A (NOKIA NETWORKS OY ;KATZ MARCOS (FI); KESKITALO ILKKA (FI); YLITALO) 18 October 2001 (2001-10-18) * page 14, line 10 - page 15, line 14 * * figure 3 *	8-45	

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The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	10 April 2003	Ghigliotti, L

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& : member of the same patent family, corresponding document

EPO FORM 1503 03.82 (P04C01)

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 02 25 4685

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
X,D	NEGI R ET AL: "Adaptive antennas for space-time coding over block-time invariant multipath fading channels" VEHICULAR TECHNOLOGY CONFERENCE, 1999 IEEE 49TH HOUSTON, TX, USA 16-20 MAY 1999, PISCATAWAY, NJ, USA,IEEE, US, 16 May 1999 (1999-05-16), pages 70-74, XP010341916 ISBN: 0-7803-5565-2 * section 3 *	8-22, 24-28	
A,D	EP 1 087 545 A (UNIV SINGAPORE) 28 March 2001 (2001-03-28) * the whole document *	8-45	
			TECHNICAL FIELDS SEARCHED (Int.Cl.7)

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	10 April 2003	Ghigliotti, L

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EPO FORM 1503 03.82 (P04C01)

European Patent Office

Application Number
EP 02 25 4685

CLAIMS INCURRING FEES

The present European patent application comprised at the time of filing more than ten claims.

☐ Only part of the claims have been paid within the prescribed time limit. The present European search report has been drawn up for the first ten claims and for those claims for which claims fees have been paid, namely claim(s):

☐ No claims fees have been paid within the prescribed time limit. The present European search report has been drawn up for the first ten claims.

LACK OF UNITY OF INVENTION

The Search Division considers that the present European patent application does not comply with the requirements of unity of invention and relates to several inventions or groups of inventions, namely:

see sheet B

☒ All further search fees have been paid within the fixed time limit. The present European search report has been drawn up for all claims.

☐ As all searchable claims could be searched without effort justifying an additional fee, the Search Division did not invite payment of any additional fee.

☐ Only part of the further search fees have been paid within the fixed time limit. The present European search report has been drawn up for those parts of the European patent application which relate to the inventions in respect of which search fees have been paid, namely claims:

☐ None of the further search fees have been paid within the fixed time limit. The present European search report has been drawn up for those parts of the European patent application which relate to the invention first mentioned in the claims, namely claims:

European Patent Office

LACK OF UNITY OF INVENTION SHEET B

Application Number
EP 02 25 4685

The Search Division considers that the present European patent application does not comply with the requirements of unity of invention and relates to several inventions or groups of inventions, namely:

1. Claims: 1-7

   Method and system for achieving diversity gain in frequency selective fading channel, by means of space-time encoding and zero-forcing pre-equalisation.

2. Claims: 8-45

   Method and system for achieving simultaneous beamforming and transmit diversity gain by means of a a space-time encoder and a transmit beamformer.

ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO. EP 02 25 4685

This annex lists the patent family members relating to the patent documents cited in the above-mentioned European search report.
The members are as contained in the European Patent Office EDP file on
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10-04-2003

Patent document cited in search report		Publication date	Patent family member(s)		Publication date
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EPO FORM P0459

For more details about this annex : see Official Journal of the European Patent Office, No. 12/82

(19) Europäisches Patentamt
European Patent Office
Office européen des brevets

(11) EP 1 387 545 A2

(12) **EUROPEAN PATENT APPLICATION**

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04.02.2004 Bulletin 2004/06

(51) Int Cl.[7]: **H04L 27/26**

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Designated Extension States:
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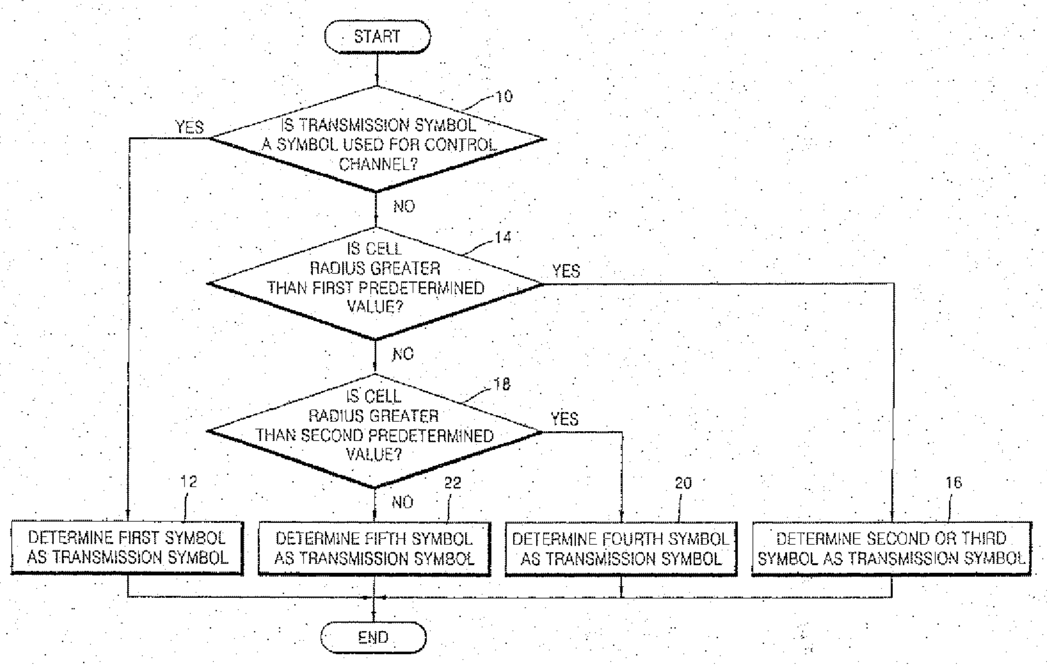
(54) Multicarrier transmission with adaptation to channel characteristics

(57) An orthogonal frequency division multiplexing (OFDM) communication method and apparatus adapted to channel characteristics are provided. The OFDM communication method includes changing at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol depending on a type of the transmission symbol and a radius of a cell, in which communication is performed. The OFDM communication apparatus includes a symbol inspector, for inspecting a type of a transmission symbol and outputting the result of the inspection as a first control signal, and a symbol and format converter, for changing at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol in response to the first control signal and a radius of a cell, in which communication is performed.

FIG. 1



EP 1 387 545 A2

Printed by Jouve, 75001 PARIS (FR)

Description

[0001] The present invention relates to orthogonal frequency division multiplexing (OFDM) communication, and more particularly, to an OFDM communication method and apparatus adapted to channel characteristics.

[0002] With a variety of environments in which a communication method is used, the communication method is required to be effective even if Doppler frequency or delay spread changes. However, since an optimum physical layer varies with channel change speed and delay spread, it is difficult to efficiently support a communication method using a single physical layer. Accordingly, a hierarchical cell including a variety of cells is used in a single communication method.

[0003] When using such a hierarchical cell, channels for users corresponding to different layers have different characteristics. For example, when a cell has a large radius, delay spread is long, and a channel change speed is fast. Accordingly, if the same modulation method is applied to different layers, a communication method cannot be adapted to the channel characteristics. In order to overcome this problem, a conventional communication method uses OFDM when the channel change speed is slow and uses code division multiple access (CDMA) when the channel change speed is fast. As described above, when using the conventional communication method, two modems of different types need to be provided for a terminal. Accordingly, the conventional communication method increases the complexity of transmitter and receiver of a terminal. In addition, since signals having different spectrum characteristics are used, the conventional communication method is difficult to develop, and radio resource management such as handover and association is difficult.

[0004] According to an aspect of the present invention, there is provided an OFDM communication method adapted to channel characteristics, including changing at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol depending on a type of the transmission symbol and a radius of a cell, in which communication is performed.

[0005] The invention thus provides an orthogonal frequency division multiplexing (OFDM) communication method through which at least one of the length of a transmission symbol, the format of a transmission symbol, and the format of a frame is changed to adapt to channel characteristics such as channel change speed and channel spread.

[0006] According to another aspect of the present invention, there is provided an OFDM communication apparatus adapted to channel characteristics, including a symbol inspector, which inspects a type of a transmission symbol and outputs the result of the inspection as a first control signal; and a symbol and format converter, which changes at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol in response to the first control signal and a radius of a cell, in which communication is performed.

[0007] The invention thus also provides an OFDM communication apparatus for performing the OFDM communication method of the invention, which is adapted to the channel characteristics.

[0008] The above and other features and advantages of the present invention will become more apparent by describing in detail preferred embodiments thereof with reference to the attached drawings in which:

FIG. 1 is a flowchart of an orthogonal frequency division multiplexing (OFDM) communication method adapted to channel characteristics according to a first embodiment of the present invention;
FIG. 2 is a diagram showing an example of a single frame including symbols having various lengths;
FIG. 3 is a flowchart of an OFDM communication method adapted to channel characteristics according to a second embodiment of the present invention;
FIG. 4 is a diagram showing an example of a macro format;
FIG. 5 is a diagram showing an example of a micro format;
FIG. 6 is a diagram showing an example of a pico format;
FIG. 7 is a flowchart of an embodiment of step 16 shown in FIG. 1 according to the present invention;
FIG. 8 is a diagram showing a hierarchical cell structure;
FIG. 9 is a diagram showing an example of a usual multiplex carrier wave transmission symbol;
FIG. 10 is a diagram showing another example of a usual multiplex carrier wave transmission symbol;
FIG. 11 is a diagram showing still another example of a usual multiplex carrier wave transmission symbol;
FIG. 12 is a block diagram of an OFDM communication apparatus for performing an OFDM communication method of the present invention, according to an embodiment of the present invention;
FIG. 13 is a block diagram of an embodiment of a symbol and format converter shown in FIG. 12;
FIG. 14 is a block diagram of another embodiment of the symbol and format converter shown in FIG. 12;
FIG. 15 is a block diagram of a first converter shown in FIG. 13;
FIG. 16 is a graph showing changes in a bit error rate with respect to changes in Doppler frequency; and
FIG. 17 is a graph showing changes in a bit error rate with respect to changes in the number of carrier waves.

[0009] Hereinafter, preferred embodiments of an orthogonal frequency division multiplexing (OFDM) communication

method to adapt to channel characteristics according to the present invention will be described in detail with reference to the attached drawings. In an OFDM communication method to adapt to channel characteristics according to the present invention, at least one of the length of a transmission symbol, the format of a frame, and the format of a transmission symbol is changed depending on a type of transmission symbol and the radius of a cell, in which communication is performed.

[0010] Channel variation is usually measured in terms of Doppler frequency multiplied by the length of an OFDM symbol, denoted as fdTs (fd: Doppler frequency in Hz, Ts; symbol duration in seconds). When fdTs is less than 0.01, the effect of channel variation on the detection performance is negligible. However, when fdTs becomes greater than 0.01, the effect becomes noticeable. Generally speaking, a fast channel change speed is when fdTs is greater than 0.01, though this is not a hard and fast rule.

[0011] Likewise, the channel length is measured by the delay spread of a channel, which is the time delay incurred from when the first signal components arrive at the receiver to when the last signal components arrive at the receiver. For example, if the last signal arrives at the receiver 0.01 seconds after the first signal arrived at the receiver, the length of the channel is 0.01 seconds. A long, medium or short channel is a relative measure utilized in the industry to describe this relative channel length. For example, if the default symbol length is 0.1 msec, a 0.1-msec channel is considered to be a long channel, a 0.01-msec channel is considered to be a medium channel, and a 0.001-msec is considered to be a short channel. In other words, when the length of a channel, divided by the length of the default OFDM symbol, is more than 10%, it is considered to be a long channel.

[0012] FIG. 1 is a flowchart of an OFDM communication method to adapt to channel characteristics according to a first embodiment of the present invention. The OFDM communication method according to the first embodiment includes determining the length of a transmission symbol depending on a type of transmission symbol and a cell radius (steps 10 through 22).

[0013] FIG. 2 is a diagram showing an example of a single frame 40, in which symbols having various lengths are mixed. The single frame 40 includes first symbols 42 and 44, second symbols 50 and 52, third symbols 54, 56, 58, and 60, a fourth symbol 48, and a fifth symbol 46.

[0014] In the OFDM communication method according to the first embodiment of the present invention shown in FIG. 1, the length of a transmission symbol is changed depending on a type of transmission symbol and the radius of a cell, in which communication is performed.

[0015] More specifically, it is determined whether a transmission symbol is a symbol that is used for a control channel in step 10. If it is determined that the transmission symbol is the symbol that is used for the control channel, the first symbol 42 or 44 shown in FIG. 2 is determined as the transmission symbol in step 12. The first symbol 42 or 44 contains control information and has a length A. In other words, if it is determined that the transmission symbol is the symbol that is used for the control channel, the length of the transmission symbol is set to A. As described above, when a large amount of data is not necessary or when it is necessary to finely divide time, as in random access or control, the relatively short length A is determined as the length of a transmission symbol.

[0016] If it is determined that the transmission symbol is not the symbol that is used for the control channel, it is determined whether a cell radius is greater than a first predetermined value in step 14. If it is determined that the cell radius is greater than the first predetermined value, the second symbol 50 or 52 or the third symbol 54, 56, 58, or 60 shown in FIG. 2 is determined as the transmission symbol in step 16. The second symbol 50 or 52 has a length B and is suitable to channel characteristics, in which a channel change speed is slow and the length of a channel is long. The third symbol 54, 56, 58, or 60 has a length C and is suitable to channel characteristics, in which a channel change speed is fast and the length of a channel is long or short. In other words, if it is determined that the cell radius is greater than the first predetermined value, the length of the transmission symbol is set to B or C.

[0017] However, if it is determined that the cell radius is not greater than the first predetermined value, it is determined whether the cell radius is greater than a second predetermined value in step 18. Here, the second predetermined value is less than the first predetermined value. If it is determined that the cell radius is greater than the second predetermined value, the fourth symbol 48 shown in FIG. 2 is determined as the transmission symbol in step 20. The fourth symbol 48 has a length D and is suitable to channel characteristics, in which a channel change speed and the length of a channel are medium. In other words, if it is determined that the cell radius is not greater than the first predetermined value but greater than the second predetermined value, the length of the transmission symbol is set to D.

[0018] However, if it is determined that the cell radius is not greater than the second predetermined value, the fifth symbol 46 shown in FIG. 2 is determined as the transmission symbol in step 22. The fifth symbol 46 has a length E and is suitable to channel characteristics, in which a channel change speed is slow and the length of a channel is short. In other words, if it is determined that the cell radius is not greater than the second predetermined value, the length of the transmission symbol is set to E.

[0019] According to the present invention, the length D of the fourth symbol 48 is shorter than the length B of the second symbol 50, and each of the lengths A, C, and E of the respective first, third, and fifth symbols 42, 54, and 46 is shorter than the length D of the fourth symbol 48. In addition, according to the present invention, each of the lengths

B, C, D, and E of the respective second, third, fourth, and fifth symbols 50, 54, 48, and 46 may be an integer multiple of the length A of the first symbol 42, and each of the lengths B, C, and D of the respective second, third, and fourth symbols 50, 54, and 48 may be an integer multiple of the length E of the fifth symbol 46.

[0020] In order to change the length of a transmission symbol, as shown in FIG. 1, the present invention changes the number of carrier waves while fixing an entire signal bandwidth. The entire signal bandwidth indicates the result of dividing an interval between carrier waves by the length of a transmission symbol. For example, when increasing the number of carrier waves while fixing an entire signal bandwidth, a distance between carrier waves is long and the length of a transmission symbol increases. Conversely, when decreasing the number of carrier waves while fixing an entire signal bandwidth, a distance between carrier waves is short and the length of a transmission symbol decreases. As described above, the length of a transmission symbol can be changed by adjusting the number of carrier waves, in step 12, 16, 20 or 22.

[0021] FIG. 3 is a flowchart of an OFDM communication method to adapt to channel characteristics according to a second embodiment of the present invention. The OFDM communication method includes converting the format of a frame depending on a cell radius in steps 70 through 78.

[0022] FIG. 4 is a diagram showing an example of a macro format. A single frame 90 is composed of a single first symbol and a plurality of second symbols, and a plurality of third symbols.

[0023] FIG. 5 is a diagram showing an example of a micro format. A single frame 92 is composed of a single first symbol and a plurality of fourth symbols.

[0024] FIG. 6 is a diagram showing an example of a pico format. A single frame 94 is composed of a single first symbol and a plurality of fifth symbols.

[0025] In the OFDM communication method according to the second embodiment of the present invention shown in FIG. 3, the format of a frame is converted depending on the radius of a cell, in which communication is performed.

[0026] For this operation, it is determined weather a cell radius is greater than a first predetermined value in step 70. If it is determined that the cell radius is greater than the first predetermined value, the format of a frame is converted into a macro format, as shown in FIG. 4, in step 72. Referring to FIG. 4, the macro format is composed of a single first symbol, a plurality of second symbols, and a plurality of third symbols. In other words, when a channel change speed is fast or slow and the length of a channel is long due to a large cell radius, the format of the frame is converted into the macro format shown in FIG. 4.

[0027] However, if it is determined that the cell radius is not greater than the first predetermined value, it is determined weather the cell radius is greater than a second predetermined value in step 74. The second predetermined value is smaller than the first predetermined value. If it is determined that the cell radius is greater than the second predetermined value, the format of a frame is converted into a micro format, as shown in FIG. 5, in step 76. Referring to FIG. 5, the micro format is composed of a single first symbol and a plurality of fourth symbols. In other words, when a channel change speed and the length of a channel are medium, the format of the frame is converted into the micro format.

[0028] However, if it is determined that the cell radius is not greater than the second predetermined value, the format of a frame is converted into a pico format, as shown in FIG. 6, in step 78. Referring to FIG. 6, the pico format is composed of a single first symbol and a plurality of fifth symbols. In other words, when a channel change speed is slow and the length of a channel is short due to a small cell radius, the format of the frame is converted into the pico format.

[0029] FIG. 7 is a flowchart of an embodiment of step 16 shown in FIG. 1 according to the present invention. The embodiment of step 16 includes determining a second or third symbol as a transmission symbol depending on a channel change speed in steps 110 through 114.

[0030] Referring to FIG. 7, if it is determined that the cell radius is greater than the first predetermined value (step 14 of FIG. 1), it is determined whether a channel change speed is greater than a predetermined speed in step 110.

[0031] If it is determined that the channel change speed is not greater than the predetermined speed, the second symbol 50 shown in FIG. 2 is determined as the transmission symbol in step 112. In other words, the length of the transmission symbol is set to B. However, if it is determined that the channel change speed is greater than the predetermined speed, the third symbol 54 is determined as the transmission symbol in step 114. In other words, the length of the transmission symbol is set to C.

[0032] According to a third embodiment of the present invention, the length of a transmission symbol and the format of a frame are changed depending on a type of transmission symbol and a cell radius. For this operation, referring to FIG. 1, if it is determined that the cell radius is greater than the first predetermined value, the second or third symbol is determined as the transmission symbol, and simultaneously the format of the frame is converted into the macro format shown in FIG. 4, in step 16. However, if it is determined that the cell radius is not greater than the first predetermined value but is greater than the second predetermined value, the fourth symbol is determined as the transmission symbol, and simultaneously the format of the frame is converted into the micro format shown in FIG. 5, in step 20. In addition, if it is determined that the cell radius is not greater than the second predetermined value, the fifth symbol is determined as the transmission symbol, and simultaneously the format of the frame is converted into the pico format shown in FIG. 6, in step 22.

[0033] FIG. 8 is a diagram showing a hierarchical cell structure, which is composed of macro cells 130, micro cells 132, and pico cells 134.

[0034] Referring to FIG. 8, the macro cells 130 represented by dotted lines correspond to cells having a radius that is greater than the first predetermined value. The micro cells 132 represented by bold sold lines correspond to cells having a radius that is not greater than the first predetermined value but greater than the second predetermined value. The pico cells 134 represented by thin solid lines correspond to cells having a radius that is not greater than the second predetermined value. The hierarchical cell structure shown in FIG. 8 is used in order to increase frequency efficiency when frequency resources are limited. As shown in FIG. 8, a plurality of micro cells 132 exist within each macro cell 130, and a plurality of pico cells 134 exist within each micro cell 132. Usually, the hierarchical cell structure is designed such that users with a fast channel change speed are gathered at the macro cells 130 and users with a slow channel change speed are gathered in the micro cells 132 or the pico cells 134. This is disclosed in pages 301-304 of a book entitled "Radio Resource Management for Wireless Networks", written by Jens Zander and Seong-Lyun Kim, and published by Artech Houser in 2001.

[0035] FIG.s 9 to 11 show examples of the multiplex carrier wave transmission symbol. FIG. 9 is an example of the second symbol and FIG.s 10 and 11 are different examples of the third symbol. In the following description, the different parts of the symbols of different examples are each given different names to avoid confusion. Therefore, the existence of a "third cyclic prefix" (for example) in a symbol should not be understood as requiring a "first" or "second" cyclic prefix in that symbol.

[0036] FIG. 9 is a diagram showing an example of a multiplex carrier wave transmission symbol. In this example, the transmission symbol is composed of a first cyclic prefix (CP) 150, a first transmission signal 158, and a first cyclic suffix (CS) 154.

[0037] According to the fourth embodiment of the present invention, the format of a symbol as well as the length of the symbol can be changed depending on a cell radius and a channel change speed.

[0038] For example, if it is determined that the channel change speed is not greater than the predetermined speed, the second symbol is determined as the transmission symbol and the format of the second symbol is converted into a format shown in FIG. 9 in step 112 of FIG. 7. In FIG. 9, the first CP 150 of the transmission symbol is the result of copying an end portion 152 of the first transmission signal 158 to the front of the first transmission signal 158 and is used to eliminate the interference of a previous symbol. The first CS 154 of the transmission symbol is the result of copying a beginning portion 156 of the first transmission signal 158 to the back of the first transmission signal 158 and is used to mitigate the alignment condition of transmission time when a carrier wave is divided and used by multiple users usually in an upward channel. Here, the first transmission signal 158 contains transmission data. As described above, since the end portion 152 of the transmission data 158 is copied to the first CP 150 and the beginning portion 156 of the transmission data 158 is copied to the first CS 154, the transmission symbol shown in FIG. 9 has a cyclic structure.

[0039] FIG. 10 is a diagram showing another example of a multiplex carrier wave transmission symbol. In this example, the transmission symbol is composed of a second CP 170, second and third transmission signals 172 and 174, and a second CS 176.

[0040] FIG. 11 is a diagram showing still another example of a multiplex carrier wave transmission symbol. In this example, the transmission symbol is composed of a third CP 190, fourth and fifth transmission signals 192 and 194, and a third CS 196.

[0041] However, if it is determined that the channel change speed is greater than the predetermined speed, the third symbol is determined as the transmission symbol and the format of the third symbol is converted into a format shown in FIG. 10 or 11 in step 114 of FIG. 7.

[0042] According to the present invention, the second CP 170 of the third symbol shown in FIG. 10 includes the end portion of transmission data stored in each of the second and third transmission signals 172 and 174 and the beginning portion of the transmission data. In other words, the second CP 170 is composed of two first CPs 150 and one first CS 154 shown in FIG. 9. In addition, each of the second and third transmission signals 172 and 174 shown in FIG. 10 contains the same transmission data as that contained in the first transmission signal 158 shown in FIG. 9. Unlike the transmission symbol shown in FIG. 9, the transmission symbol shown in FIG. 10 includes repeated transmission data following the second CP 170. Here, the second CS 176 includes the beginning portion of the transmission data. In other words, the second CS 176 is composed of one first CS 154 shown in FIG. 9.

[0043] According to the present invention, the third CP 190 of the third symbol shown in FIG. 11 includes the end portions of transmission data stored in each of the fourth and fifth transmission signals 192 and 194. In other words, the third CP 190 is composed of only two first CPs 150. In addition, each of the fourth and fifth transmission signals 192 and 194 shown in FIG. 11 contains the same transmission data as that contained in the first transmission signal 158 shown in FIG. 9. Unlike the transmission symbol shown in FIG. 9, the transmission symbol shown in FIG. 11 includes repeated transmission data following the third CP 190. Here, the third CS 196 includes the beginning portions of the transmission data. In other words, the third CS 196 is composed of two first CSs 154 shown in FIG. 9. The

transmission symbol shown in FIG. 11 can be used when timing does not agree well as a whole as in random access.

[0044] Consequently, in an OFDM communication method, the length of the first CP 150 is required to be longer than a channel length. However, since duplicate information is contained in the first CP 150, communication efficiency is decreased when the result of dividing the length of the first CP 150 by the transmission data contained in the first transmission signal 158 is too large. Accordingly, in order to maintain the communication efficiency, the length of the transmission data needs to be 5-10 times longer than the length of the first CP 150. In other words, in the transmission symbol, the length of the first CP 150 needs to be as short as possible. Here, if the length of the transmission symbol is long in a state in which a channel change speed is fast, a channel may change within the transmission data, and thus communication performance may be degraded. As described above, it is necessary to increase the length of the first CP 150 as a channel length increases, but there is a limitation in increasing the length of the first CP 150. In order to solve this problem, an OFDM communication method according to the present invention described above adaptively changes the length of a transmission symbol depending on a channel change speed and a cell radius, as shown in FIG. 1. As described above, influence of inter symbol interference (ISI) can be overcome while the degree of overhead due to the first CP 150 is fixed, by adaptively changing the length of a transmission symbol.

[0045] Hereinafter, the structure and operation of an OFDM communication apparatus adapted to channel characteristics according to the present invention, which performs the above-described OFDM communication method adapted to channel characteristics according to the present invention, will be described with reference to the attached drawings.

[0046] FIG. 12 is a block diagram of an OFDM communication apparatus for performing the above-described OFDM communication method according to an embodiment of the present invention. The OFDM communication apparatus includes a symbol inspector 210 and a symbol and format converter 212.

[0047] Referring to FIG. 12, in order to perform step 10 shown in FIG. 1, the symbol inspector 210 inspects the type of a transmission symbol that is input through an input terminal IN1 and outputs the result of the inspection to the symbol and format converter 212 as a first control signal C1. For example, the symbol inspector 210 inspects whether the transmission symbol input through the input terminal IN1 is a symbol used for a control channel and outputs the result of the inspection as the first control signal C1.

[0048] In order to perform steps 12 through 22 shown in FIG. 1, the symbol and format converter 212 changes at least one of the length of the transmission symbol, the format of a frame, and the format of the transmission symbol in response to a cell radius that is input through an input terminal IN2 and the first control signal C1 received from the symbol inspector 210, and outputs the result of the change through an output terminal OUT1. Here, what will be changed among the length of the transmission symbol, the format of a frame, and the format of the transmission symbol is predetermined.

[0049] The following description concerns the structures and operations of embodiments of the symbol and format converter 212 shown in FIG. 12 according to the present invention. FIG.s 13 and 14 each show an example of the converter 212. In the following description, the different parts of the converter in the two examples are each given different names to avoid confusion. Therefore, the existence of a "third comparator" (for example) in the converter should not be understood as requiring a "first" or "second" comparator in that converter.

[0050] FIG. 13 is a block diagram of an embodiment 212A of the symbol and format converter 212 shown in FIG. 12. The embodiment 212A includes a first comparator 230, a second comparator 232, and a first converter 234.

[0051] In order to perform step 14 shown in FIG. 1, the first comparator 230 of the symbol and format converter 212A shown in FIG. 13 compares the cell radius that is input through an input terminal IN3 with a first predetermined value in response to the first control signal C1 that is input from the symbol inspector 210 and outputs the result of the comparison to the second comparator 232 and the first converter 234 as a second control signal C2. In other words, when it is recognized based on the first control signal C1 received from the symbol inspector 210 that the transmission symbol is not a symbol used for a control channel, the first comparator 230 compares the cell radius with the first predetermined value. Here, the first predetermined value may be set in the first comparator 230 in advance, as shown in FIG. 13, or may be externally input, unlike the structure shown in FIG. 13.

[0052] In order to perform step 18 shown in FIG. 1, the second comparator 232 compares the cell radius with a second predetermined value in response to the second control signal C2 received from the first comparator 230 and outputs the result of the comparison to the first converter 234 as a third control signal C3. For example, when it is recognized based on the second control signal C2 received from the first comparator 230 that the cell radius is not greater than the first predetermined value, the second comparator 232 compares the cell radius with the second predetermined value and outputs the result of the comparison as the third control signal C3. Here, the second predetermined value may be set in the second comparator 232 in advance, as shown in FIG. 13, or may be externally input, unlike the structure shown in FIG. 13.

[0053] In order to perform steps 12, 16, 20, and 22 shown in FIG. 1, the first converter 234 determines one among first through fifth symbols as the transmission symbol in response to the first control signal C1 received from the symbol inspector 210, the second control signal C2 received from the first comparator 230, and the third control signal C3

received from the second comparator 232 and outputs the determined symbol through an output terminal OUT2. For example, in order to perform step 12, when it is recognized based on the first control signal C1 received from the symbol inspector 210 that the transmission symbol is a symbol used for a control channel, the first converter 234 determines the length of the transmission symbol as A. However, in order to perform step 16, when it is recognized based on the first and second control signals C1 and C2 that the transmission symbol is not a symbol used for a control channel and the cell radius is greater than the first predetermined value, the first converter 234 determines the length of the transmission symbol as B or C. In addition, in order to perform step 20, when it is recognized based on the first, second, and third control signals C1, C2 and C3 that the transmission symbol is not a symbol used for a control channel and the cell radius is not greater than the first predetermined value but is greater than the second predetermined value, the first converter 234 determines the length of the transmission symbol as D. In order to perform step 22, when it is recognized based on the first, second, and third control signals C1, C2 and C3 that the transmission symbol is not a symbol used for a control channel and the cell radius is not greater than the first predetermined value and is not greater than the second predetermined value, the first converter 234 determines the length of the transmission symbol as E.

[0054] FIG. 14 is a block diagram of another embodiment 212B of the symbol and format converter 212 shown in FIG. 12. The embodiment 212B includes a third comparator 250, a fourth comparator 252, and a second converter 254.

[0055] The symbol and format converter 212B shown in FIG. 14 performs the OFDM communication method shown in FIG. 3. In order to perform step 70 shown in FIG. 3, the third comparator 250 of the symbol and format converter 212B shown in FIG. 14 compares the cell radius that is input through an input terminal IN4 with the first predetermined value and outputs the result of the comparison to the fourth comparator 252 and the second converter 254 as a fourth control signal C4. Here, the first predetermined value may be set in the third comparator 250 in advance, as shown in FIG. 14, or may be externally input, unlike the structure shown in FIG. 14.

[0056] In order to perform step 74 shown in FIG. 3, the fourth comparator 252 compares the cell radius input through the input terminal IN4 with the second predetermined value in response to the fourth control signal C4 received from the third comparator 250 and outputs the result of the comparison to the second converter 254 as a fifth control signal C5. For example, when it is recognized based on the fourth control signal C4 received from the third comparator 250 that the cell radius is not grater than the first predetermined value, the fourth comparator 252 compares the cell radius with the second predetermined value and outputs the result of the comparison as the fifth control signal C5.

[0057] In order to perform steps 72, 76, and 78 of FIG. 3, the second converter 254 converts the format of a frame into a macro format, micro format, or pico format in response to the fourth control signal C4 received from the third comparator 250 and the fifth control signal C5 received from the fourth comparator 252 and outputs the frame having the converted format through an output terminal OUT3. For example, in order to perform step 72, when it is recognized based on the fourth control signal C4 that the cell radius is grater than the first predetermined value, the second converter 254 converts the format of a frame into the macro format shown in FIG. 4. In order to perform step 76, when it is recognized based on the fourth and fifth control signals C4 and C5 that the cell radius is not grater than the first predetermined value but is greater than the second predetermined value, the second converter 254 converts the format of a frame into the micro format shown in FIG. 5. In order to perform step 78, when it is recognized based on the fourth and fifth control signals C4 and C5 that the cell radius is less than both first and second predetermined values, the second converter 254 converts the format of a frame into the pico format shown in FIG. 6.

[0058] According to an embodiment of the present invention, the symbol and format converter 212 shown in FIG. 12 may be provided with the symbol and format converter 212A shown in FIG. 13 in order to perform steps 12 through 22 shown in FIG. 1 and the symbol and format converter 212B shown in FIG. 14 in order to perform the OFDM communication method shown in FIG. 3.

[0059] According to another embodiment of the present invention, the symbol and format converter 212 shown in FIG. 12 may be provided with only the symbol and format converter 212A shown in FIG. 13 in order to perform steps 12 through 22 shown in FIG. 1 and the OFDM communication method shown in FIG. 3. In this situation, the symbol and format converter 212A shown in FIG. 13 can perform all of the steps 12 through 22 shown in FIG. 1 and the OFDM communication method shown in FIG. 3. For example, the first and second comparators 230 and 232 perform steps 14 and 18, respectively, shown in FIG. 1 and also perform steps 70 and 74, respectively, shown in FIG. 3. The first converter 234 performs steps 12, 16, 20, 22, 72, 76, and 78. In other words, the first converter 234 converts the format of a frame into a macro, micro, or pico format in response to the second and third control signals C2 and C3 respectively received from the first and second comparators 230 and 232 and outputs the frame having the converted format through the output terminal OUT2. For example, in order to perform step 72, when it is recognized based on the second control signal C2 that the cell radius is greater than the first predetermined value, the first converter 234 converts the format of a frame into the macro format shown in FIG. 4. In order to perform step 76, when it is recognized based on the second and third control signals C2 and C3 that the cell radius is not greater than the first predetermined value but is greater than the second predetermined value, the first converter 234 converts the format of a frame into the micro format shown in FIG. 5. In order to perform step 78, when it is recognized based on the second and third control signals C2 and C3 that the cell radius is not greater than both first and second predetermined values, the first converter 234

converts the format of a frame into the pico format shown in FIG. 6.

[0060] FIG. 15 is a block diagram of the first converter 234 shown in FIG. 13. The first converter 234 includes a fifth comparator 270 and a format converter 272.

[0061] The first converter 234 shown in FIG. 13 may include the fifth comparator 270 in order to perform step 110 shown in FIG. 7. In this situation, the fifth comparator 270 compares a channel change speed with a predetermined speed in response to the second control signal C2 received from the first comparator 230 and outputs the result of the comparison as a sixth control signal C6. For example, when it is recognized based on the second control signal C2 that the cell radius is greater than the first predetermined value, the fifth comparator 270 compares the channel change speed with the predetermined speed and outputs the result of the comparison as the sixth control signal C6. Here, the format converter 272 of the first converter 234 converts the format of the determined transmission symbol in response to the sixth control signal C6 received from the fifth comparator 270 and outputs the transmission symbol having the converted format through the output terminal OUT2. For example, when it is recognized based on the sixth control signal C6 received from the fifth comparator 270 that the channel change speed is not greater than the predetermined speed, the format converter 272 converts the format of the transmission symbol into the format shown in FIG. 9 in order to perform step 112. However, when it is recognized based on the sixth control signal C6 received from the fifth comparator 270 that the channel change speed is greater than the predetermined speed, the format converter 272 converts the format of the transmission symbol into the format shown in FIG. 10 or 11 in order to perform step 114.

[0062] When an OFDM communication method and apparatus adapted to channel characteristics according to the present invention are used at a whole signal bandwidth of 20 MHz, the results of operation are obtained, as shown in Table 1.

Table 1

Division	First symbol	Second symbol	Third symbol	Fourth symbol	Fifth symbol
Number of carrier waves	512	4096	1024	2048	1024
Ts	0.02844	0.2275	0.05689	0.1138	0.05689
Tg	2.81	22.45	5.6	11.22	5.6
Lamp (Up) (μs)	1	1	1	1	1
CS	1	1	1	1	1
CP	0.81	20.45	3.6	9.22	3.6
Ts+Tg	0.0313	0.25	0.0625	0.125	0.0625
Bit rate	4 Mbps	8-50 Mbps	4-25 Mbps	8-50 Mbps	8-50 Mbps

[0063] Here, Ts denotes a period of time indicating the length of a transmission symbol, Tg denotes a guard time, the unit of the CS is in μs, and bps indicates bits per second.

[0064] As is seen from Table 1, in an OFDM communication method and apparatus according to the present invention, the length of a transmission symbol Ts is adjusted by changing the number of carrier waves so that the method and apparatus can be adapted to various communication environments. The transmission symbol is adjusted to adapt to various communication environments for the following reasons.

[0065] For example, let's assume that a Veh B channel, from Tr 101 146 v3.0, which is disclosed in a book entitled "Digital Communications", written by J. Proakis, and published by McGraw Hill in 1995, is used; the number of carrier waves is 4096; a whole signal bandwidth is 18 MHz; and a spread factor (SF) is 4.

[0066] FIG. 16 is a graph showing changes in a bit error rate (BER) with respect to changes in Doppler frequency. The vertical axis indicates a BER, and the horizontal axis indicates Eb/No where Eb is energy per bit and No is the variance of noise.

[0067] The BER at a Doppler frequency, i.e., a channel change speed, of 170 (■) is greater than the BER at a channel change speed of 17 (★). The BER at a channel change speed of 500 (▲) is greater than the BER at the channel change speed of 170 (■). Consequently, as is seen from FIG. 16, the BER increases with an increase in a channel change speed.

[0068] In the meantime, when the same assumption as described above is adopted, with the exception that the SF is 1, the Doppler frequency is 500, and a Veh A channel is used instead of the Veh B channel, changes in a BER with respect to changes in the number of carrier waves will be described below. Here, the channels (Veh A and Veh B) are disclosed in Table 1.2.2.3 in page 43 of a book entitled "Selection Procedures for the Choice of Radio Transmission Technologies" and published by Universal Mobile Telecommunication System (UMTS), which is under a standardization group of European Telecommunications Standardization Institute (ETSI), in Technical Report (TR) 101112 of the ETSI.

[0069] FIG. 17 is a graph showing changes in a BER with respect to changes in the number of carrier waves. The vertical axis indicates a BER, and the horizontal axis indicates Eb/No.

[0070] As shown in FIG. 17, the BER when the number of carrier waves is 2048 (■) is greater than the BER when the number of carrier waves is 1024 (▲), and the BER when the number of carrier waves is 4096 (★) is greater than the BER when the number of carrier waves is 2048 (■). Consequently, as is seen from FIG. 17, when the length of a transmission symbol is decreased by decreasing the number of carrier waves from 4096 to 1096, the influence of the Doppler frequency is reduced, thereby decreasing the BER. Accordingly, if transmission data is repeated two times, as shown in FIG. 10 or 11, influence due to a change in a channel length is decreased and interchannel interference is prevented.

[0071] As described above, in an OFDM communication method and apparatus to adapt to channel characteristics according to the present invention, at least one of the length and the format of a transmission symbol and the format of a frame is changed to adapt to channel characteristics such as a channel change speed and a channel length so that communication can be accomplished at a low BER and high efficiency under various environments and a terminal can be simply implemented. In particular, under an environment in which a channel change speed is fast and a channel length is long, communication reliability can be enhanced. Since the transmission symbol includes the first symbol regardless of a cell radius, as shown in FIGS. 4 through 6, the present invention facilitates wireless resource management such as association and handover.

[0072] While the present invention has been particularly shown and described with reference to exemplary embodiments thereof, it will be understood by those of ordinary skill in the art that various changes in form and details may be therein without departing from the scope of the invention as defined by the appended claims.

Claims

1. An orthogonal frequency division multiplexing (OFDM) communication method to adapt to channel characteristics, comprising the steps of changing at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol depending on a type of the transmission symbol and a radius of a cell, in which communication is performed.

2. The OFDM communication method of claim 1, wherein the changing steps comprises the following steps:

   (a) determining whether the transmission symbol is a symbol that is used for a control channel;
   (b) if it is determined that the transmission symbol is the symbol that is used for a control channel, determining a first symbol containing control information as the transmission symbol;
   (c) if it is determined that the transmission symbol is not the symbol that is used for a control channel, determining whether the cell radius is greater than a first predetermined value;
   (d) if it is determined that the cell radius is greater than the first predetermined value, determining a second symbol, which is suitable to channel characteristics where a channel change speed is slow and a channel length is long, or a third symbol, which is suitable to channel characteristics where the channel change speed is fast and the channel length is long, as the transmission symbol;
   (e) if it is determined that the cell radius is not greater than the first predetermined value, determining whether the cell radius is greater than a second predetermined value;
   (f) if it is determined that the cell radius is greater than the second predetermined value, determining a fourth symbol, which is suitable to channel characteristics where the channel change speed and the channel length are medium, as the transmission symbol; and
   (g) if it is determined that the cell radius is not greater than the second predetermined value, determining a fifth symbol, which is suitable to channel characteristics where the channel change speed is slow and the channel length is short, as the transmission symbol,

   wherein the second predetermined value is less than the first predetermined value, a length of the fourth symbol is less than a length of the second symbol, and a length of each of the first, third, and fifth symbols is less than the length of the fourth symbol.

3. The OFDM communication method of claim 2, wherein the length of each of the second, third, fourth, and fifth symbols is an integer multiple of the length of the first symbol.

4. The OFDM communication method of claim 2, wherein the length of each of the second, third, and fourth symbols is an integer multiple of the length of the fifth symbol.

5. The OFDM communication method of any one of claims 2 to 4, further comprising the step of adjusting the length of the determined transmission symbol by changing the number of carrier waves

6. The OFDM communication method of claim 2, wherein step (d) comprises determining the second or third symbol as the transmission symbol and converting the format of the frame into a macro format if it is determined that the cell radius is greater than the first predetermined value,

   step (f) comprises determining the fourth symbol as the transmission symbol and converting the format of the frame into a micro format if it is determined that the cell radius is greater than the second predetermined value, and

   step (g) comprises determining the fifth symbol as the transmission symbol and converting the format of the frame into a pico format if it is determined that the cell radius is not greater than the second predetermined value.

7. The OFDM communication method of claim 1, wherein changing step comprises the steps of:

   (h) determining whether the radius cell is greater than a first predetermined value;
   (i) if it is determined that the radius cell is greater than the first predetermined value, converting the format of the frame into a macro format;
   (j) if it is determined that the radius cell is not greater than the first predetermined value, determining whether the radius cell is greater than a second predetermined value;
   (k) if it is determined that the radius cell is greater than the second predetermined value, converting the format of the frame into a micro format; and
   (l) if it is determined that the radius cell is not greater than the second predetermined value, converting the format of the frame into a pico format,

   wherein the first predetermined value is greater than the second predetermined value.

8. The OFDM communication method of claim 7, wherein the macro format comprises: a first symbol, which contains control information;

   a second symbol, which is suitable to channel characteristics where a channel change speed is slow and a channel length is long; and

   a third symbol, which is suitable to channel characteristics where the channel change speed is fast and the channel length is long.

9. The OFDM communication method of claim 7 or 8, wherein the micro format comprises:

   a first symbol, which contains control information; and
   a fourth symbol, which is suitable to channel characteristics where a channel change speed and a channel length are medium.

10. The OFDM communication method of claim 7, 8 or 9, wherein the pico format comprises:

    a first symbol, which contains control information; and
    a fifth symbol, which is suitable to channel characteristics where a channel change speed is slow and a channel length is short.

11. The OFDM communication method of claim 2, wherein step (d) further comprises the steps of:

    (d1) if it is determined that the cell radius is greater than the first predetermined value, determining whether the channel change speed is greater than a predetermined speed;
    (d2) if it is determined that the channel change speed is not greater than the predetermined speed, determining the second symbol as the transmission symbol; and
    (d3) if it is determined that the channel change speed is greater than the predetermined speed, determining the third symbol as the transmission symbol.

12. The OFDM communication method of claim 11, wherein the second symbol determined as the transmission symbol in step (d2) comprises:

    a first cyclic prefix, which contains an end portion of transmission data;

a first transmission signal, which contains the transmission data; and
a first cyclic suffix, which contains a beginning portion of the transmission data.

13. The OFDM communication method of claim 11, wherein the third symbol determined as the transmission symbol in step (d3) comprises:

a first cyclic prefix, which contains a plurality of end portions of transmission data and a beginning portion of the transmission data;
a first transmission signal, which contains the transmission data;
a second transmission signal, which contains the transmission data; and
a first cyclic suffix, which contains the beginning portion of the transmission data.

14. The OFDM communication method of claim 11, wherein the third symbol comprises:

a first cyclic prefix, which contains a plurality of end portions of transmission data;
a first transmission signal, which contains the transmission data;
a second transmission signal, which contains the transmission data; and
a first cyclic suffix, which contains a plurality of beginning portions of the transmission data.

15. An orthogonal frequency division multiplexing (OFDM) communication apparatus to adapt to channel characteristics, comprising:

a symbol inspector, for inspecting a type of a transmission symbol and outputting the result of the inspection as a first control signal; and
a symbol and format converter, for changing at least one of a length of a transmission symbol, a format of a frame, and a format of the transmission symbol in response to the first control signal and a radius of a cell, in which communication is performed.

16. The OFDM communication apparatus of claim 15, wherein the symbol and format converter comprises:

a first comparator, for comparing the cell radius with a first predetermined value in response to the first control signal and outputting the result of the comparison as a second control signal;
a second comparator, for comparing the cell radius with a second predetermined value in response to the second control signal and outputting the result of the comparison as a third control signal; and
a first converter, for determining one among first, second, third, fourth, and fifth symbols as the transmission symbol in response to the first, second, and third control signals and outputting the determined symbol,

wherein the second predetermined value is less than the first predetermined value, the first symbol contains control information, the second symbol is suitable to channel characteristics where a channel change speed is slow and a channel length is long, the third symbol is suitable to channel characteristics where the channel change speed is fast and the channel length is long, the fourth symbol is suitable to channel characteristics where the channel change speed and the channel length are medium, and the fifth symbol is suitable to channel characteristics where the channel change speed is slow and the channel length is short.

17. The OFDM communication apparatus of claim 15 or 16, wherein the symbol and format converter comprises:

a third comparator, for comparing the cell radius with a first predetermined value and outputting the result of the comparison as a fourth control signal;
a fourth comparator, for comparing the cell radius with a second predetermined value in response to the fourth control signal and outputting the result of the comparison as a fifth control signal; and
a second converter, for converting the format of the frame into one of a macro format, a micro format, and a pico format in response to the fourth and fifth control signals,

wherein the first predetermined value is greater than the second predetermined value.

18. The OFDM communication apparatus of claim 16, wherein the first converter converts the format of the frame into one of a macro format, a micro format, and a pico format in response to the second and third control signals.

19. The OFDM communication apparatus of claim 16, wherein the first converter comprises a fifth comparator, for comparing the channel change speed with a predetermined speed in response to the second control signal and outputting the result of the comparison as a sixth control signal, and a format converter, for converting the format of the determined symbol in response to the sixth control signal.

FIG. 1

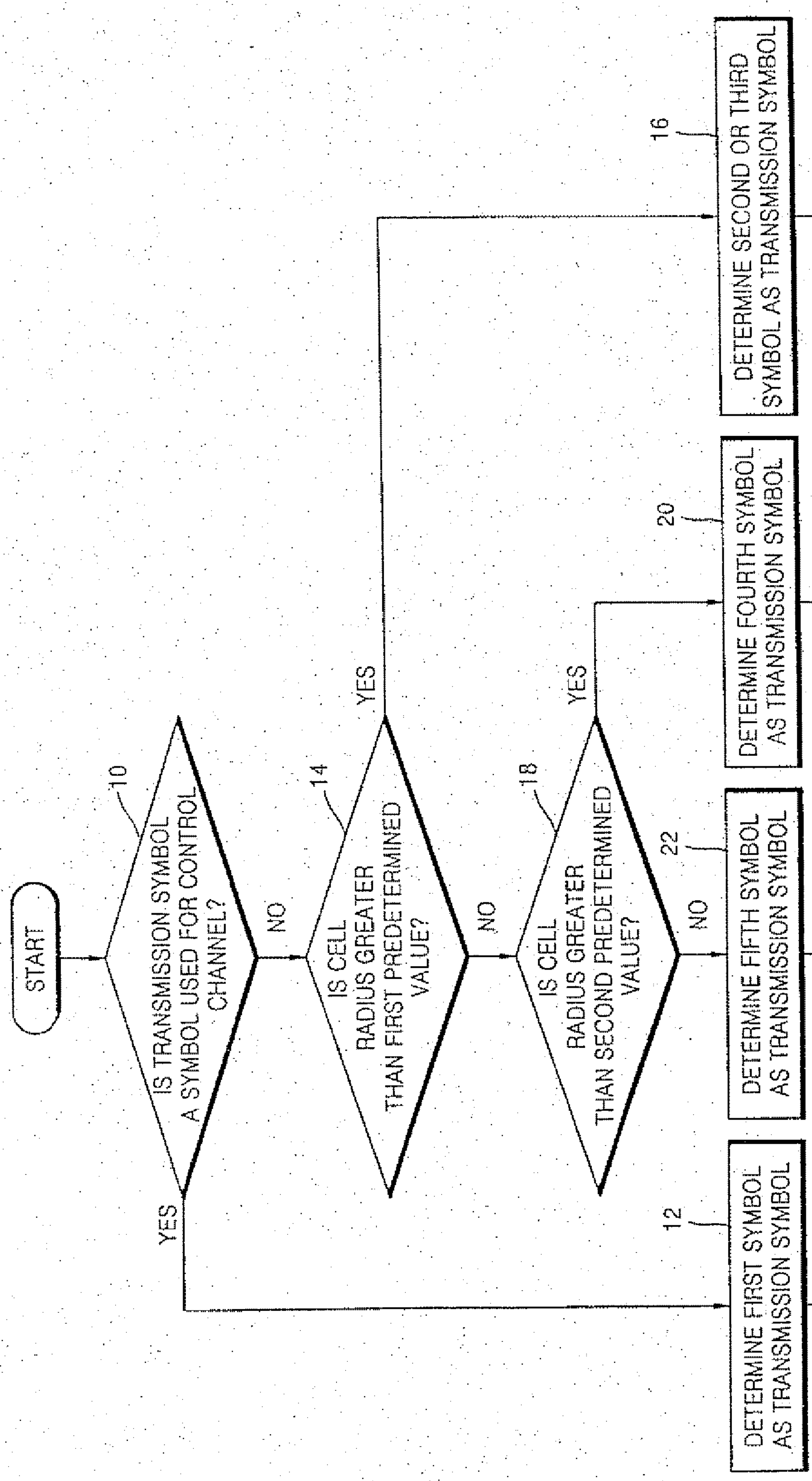

START
10
IS TRANSMISSION SYMBOL A SYMBOL USED FOR CONTROL CHANNEL?
YES
NO
12
DETERMINE FIRST SYMBOL AS TRANSMISSION SYMBOL
14
IS CELL RADIUS GREATER THAN FIRST PREDETERMINED VALUE?
YES
NO
16
DETERMINE SECOND OR THIRD SYMBOL AS TRANSMISSION SYMBOL
18
IS CELL RADIUS GREATER THAN SECOND PREDETERMINED VALUE?
YES
NO
20
DETERMINE FOURTH SYMBOL AS TRANSMISSION SYMBOL
22
DETERMINE FIFTH SYMBOL AS TRANSMISSION SYMBOL
END

FIG. 2

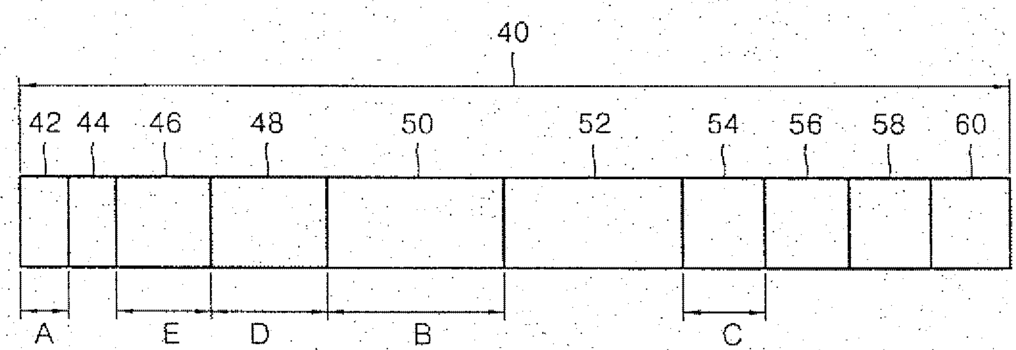


FIG. 3

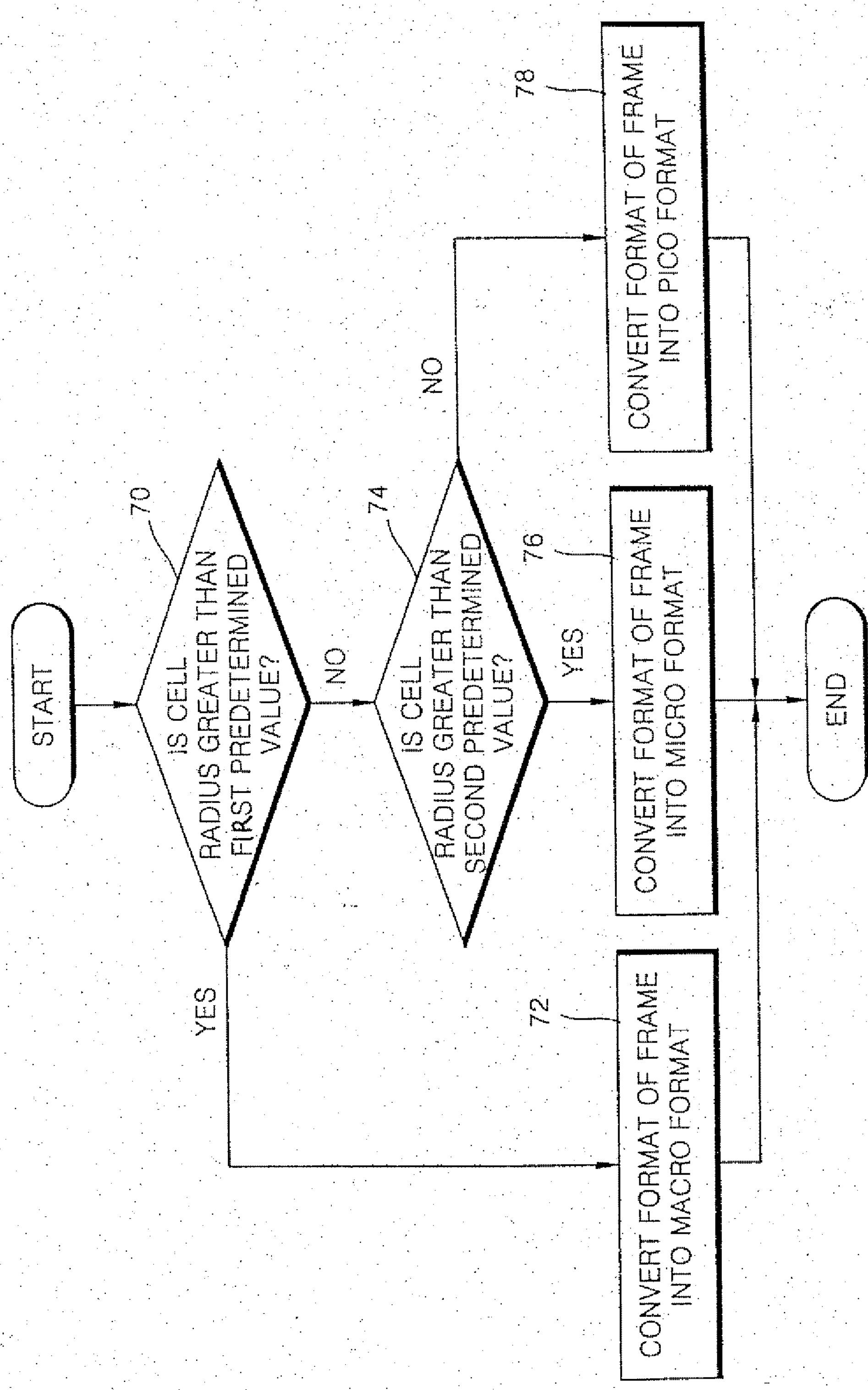

START
IS CELL RADIUS GREATER THAN FIRST PREDETERMINED VALUE?
70
YES
NO
IS CELL RADIUS GREATER THAN SECOND PREDETERMINED VALUE?
74
YES
NO
72
CONVERT FORMAT OF FRAME INTO MACRO FORMAT
76
CONVERT FORMAT OF FRAME INTO MICRO FORMAT
78
CONVERT FORMAT OF FRAME INTO PICO FORMAT
END

FIG. 4

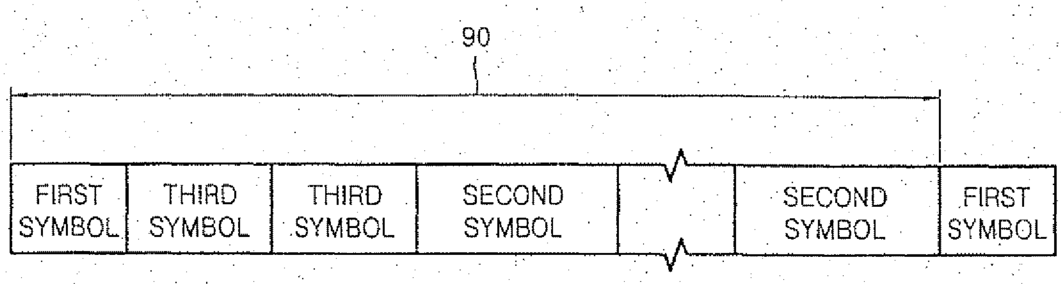


FIG. 5

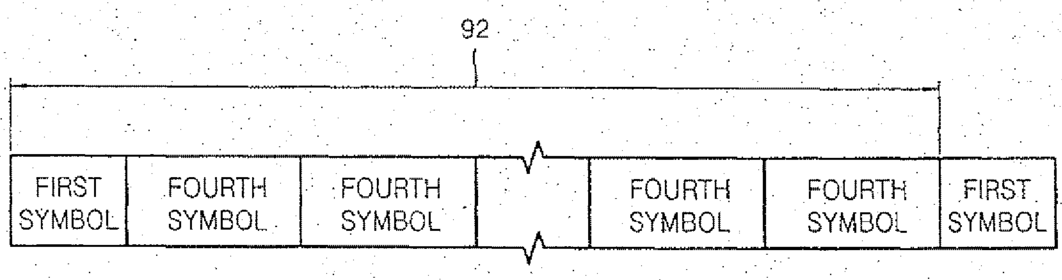


FIG. 6

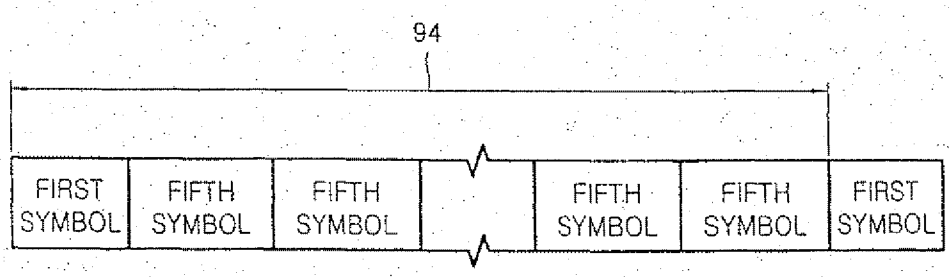


FIG. 7

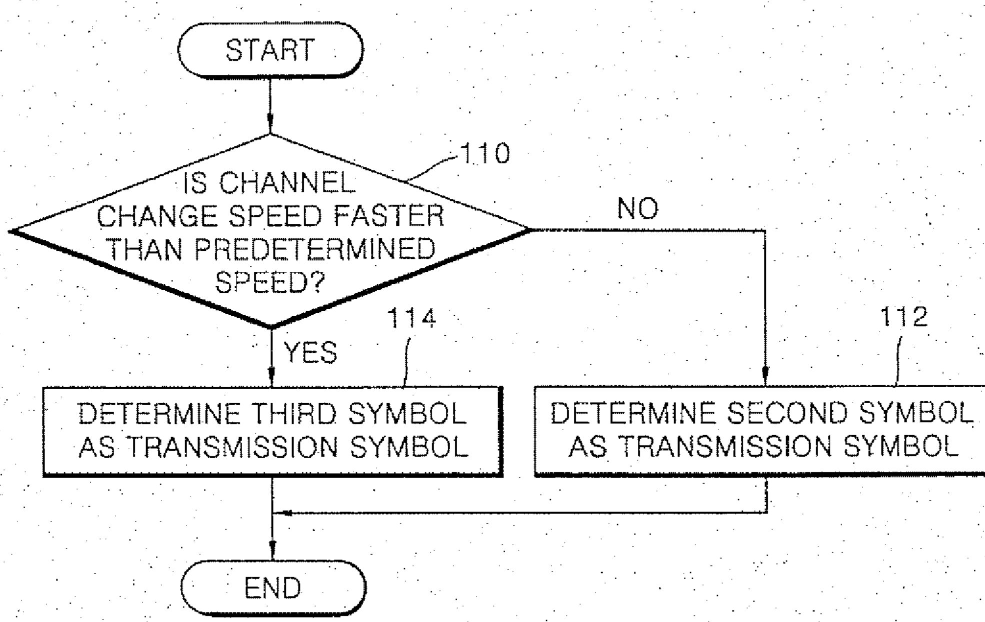

START
IS CHANNEL CHANGE SPEED FASTER THAN PREDETERMINED SPEED?
110
NO
YES
114
112
DETERMINE THIRD SYMBOL AS TRANSMISSION SYMBOL
DETERMINE SECOND SYMBOL AS TRANSMISSION SYMBOL
END

FIG. 8

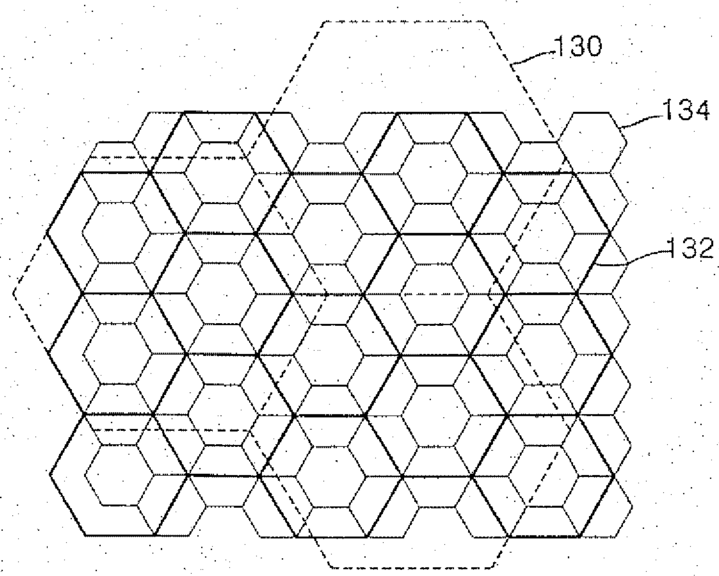

130
134
132

FIG. 9

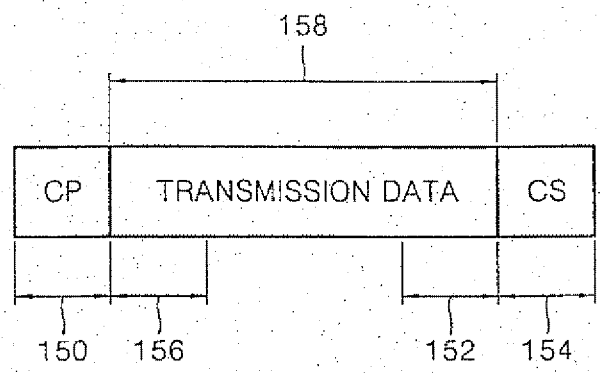


FIG. 10

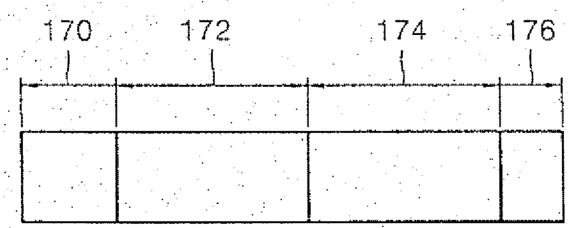


FIG. 11

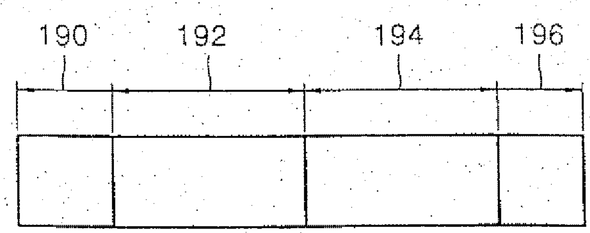


FIG. 12

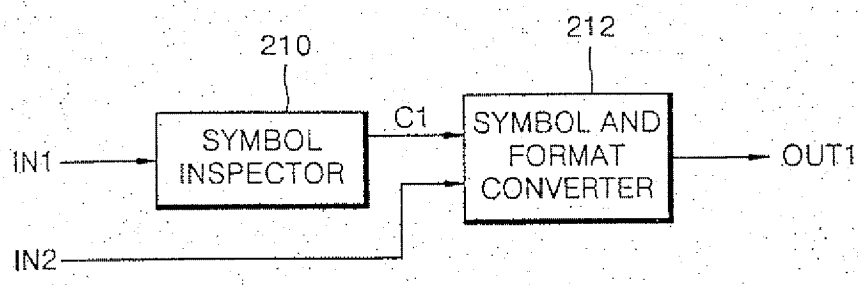


FIG. 13

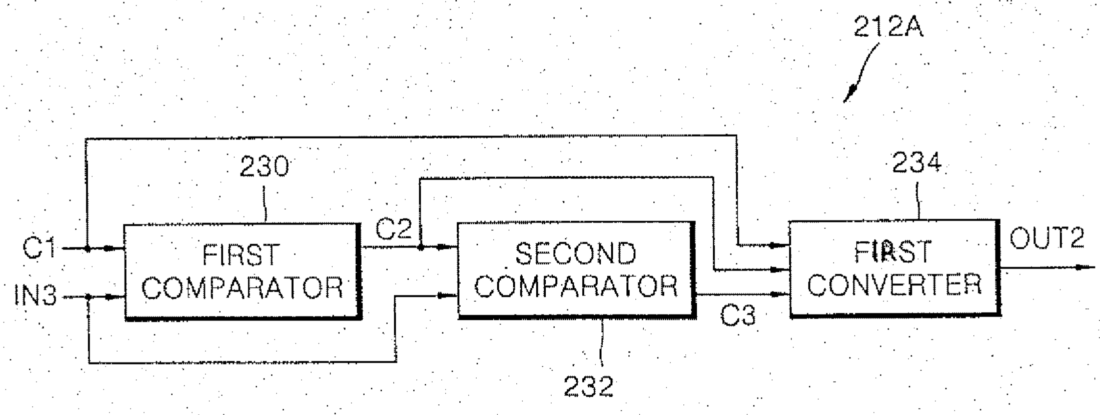


FIG. 14

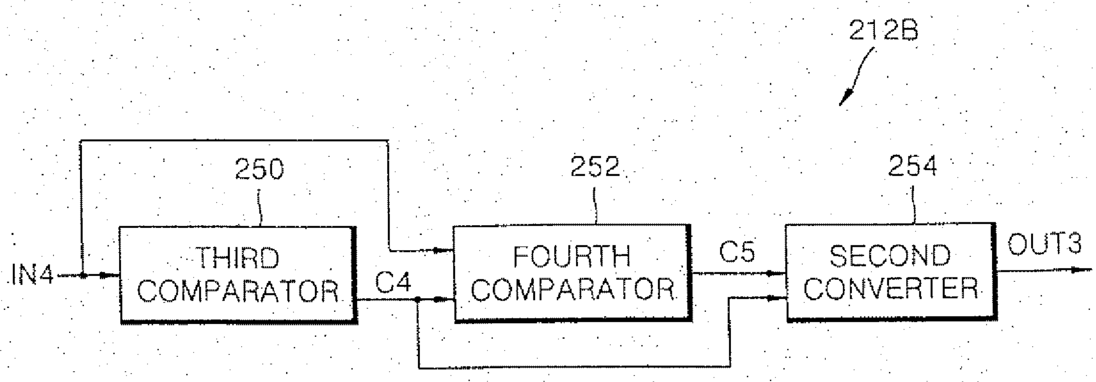


FIG. 15

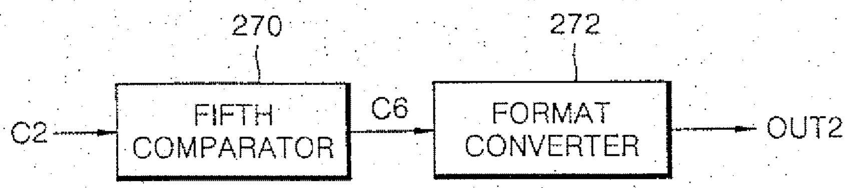


FIG. 16

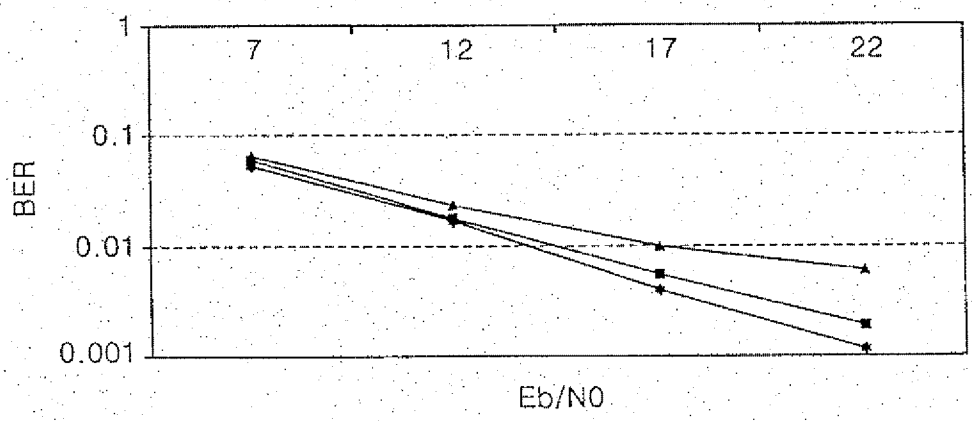
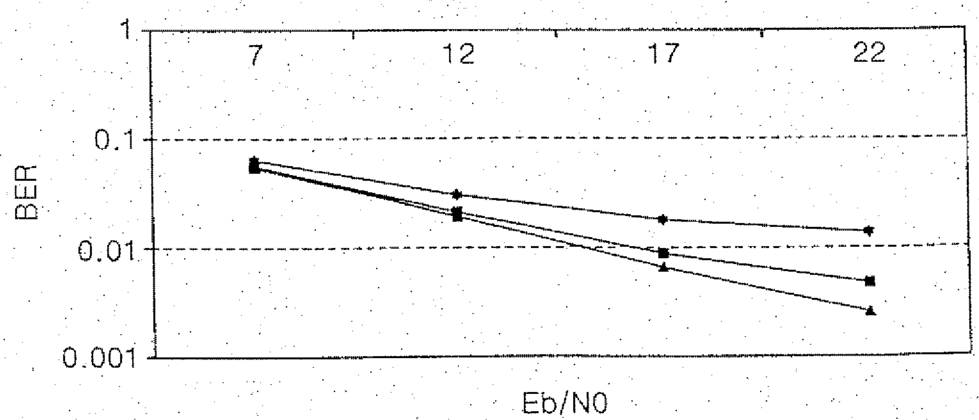


FIG. 17



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H04L 27/26

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H4P PAX PRE

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EP 0851642 A2
US 6249250 B
EP 1037303 A1
WO 1997/040608 A1

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INT CL[7] H04L 27/26
Online: EPODOC, JAPIO, WPI

(54) Abstract Title

Adaptive OFDM receiver where sets of sub-carriers which are coherent are grouped into sub-bands and a single weight is calculated for each sub-band

(57) The present invention relates to an improved adaptive weighting system particular for use with broadband multicarrier systems. Prior art systems determine a series of weights to be applied to each sub-channel of a frequency domain received signal. These weights are then applied to each of the channels according to their coherency. The present invention provides a reduction in the complexity of determining the weights by grouping the sub-channels into groups/sub-bands which are sufficiently coherent to allow narrowband processing and for a single set of weights to be applied to all sub-channels in the sub-band. Thereby reducing the processing required.

The system may use multiple antennas.

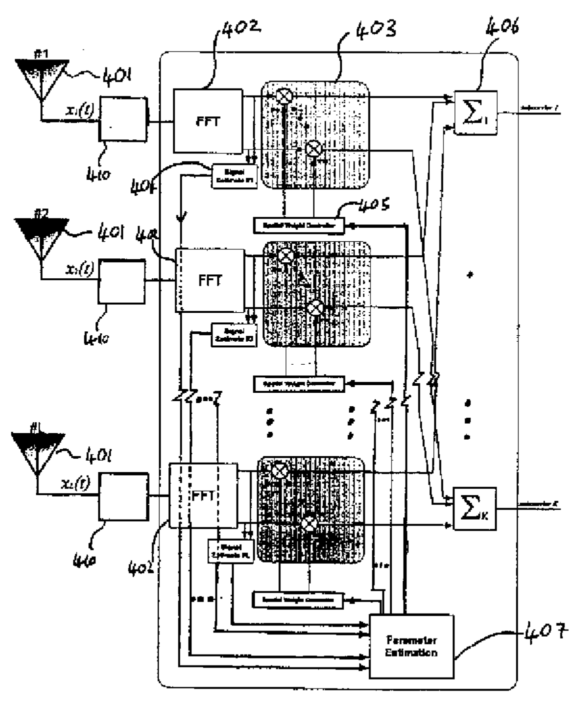


Fig. 4

GB 2 373 973 A

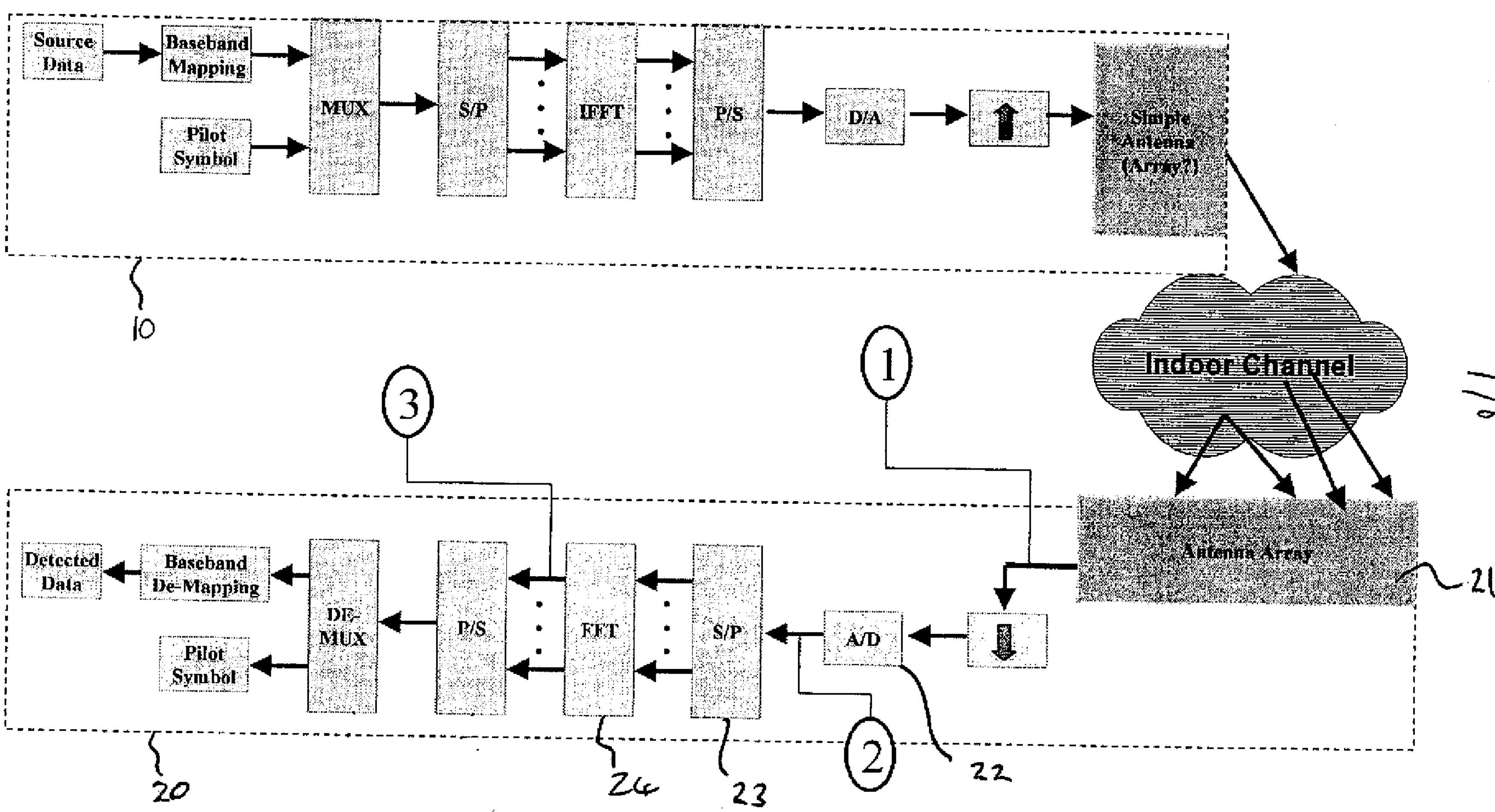
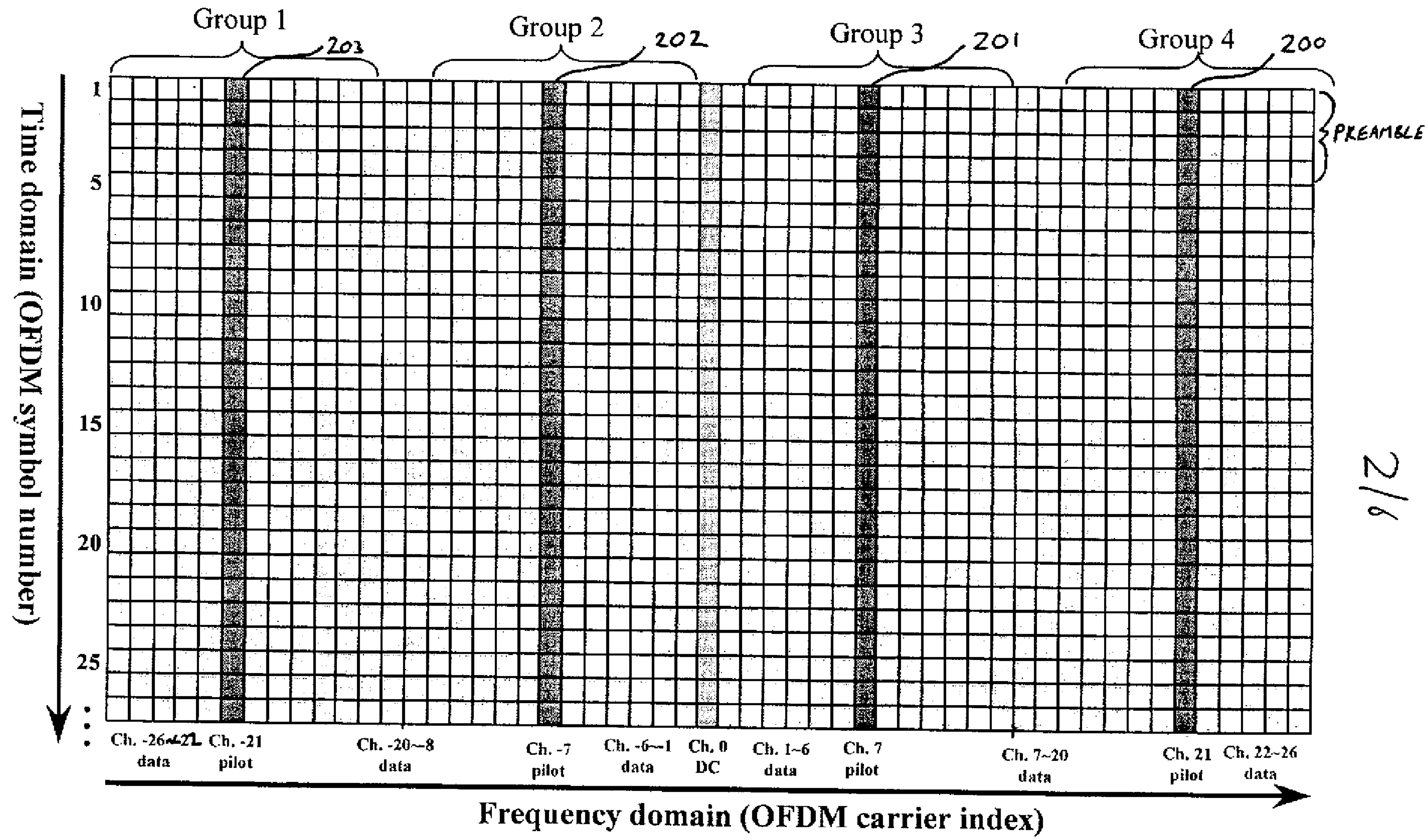

Source Data
Baseband Mapping
Pilot Symbol
MUX
S/P
IFFT
P/S
D/A
Simple Antenna (Array?)
10
Indoor Channel
Antenna Array
21
1
2
3
A/D
22
S/P
23
FFT
24
P/S
DE-MUX
Baseband De-Mapping
Pilot Symbol
Detected Data
20

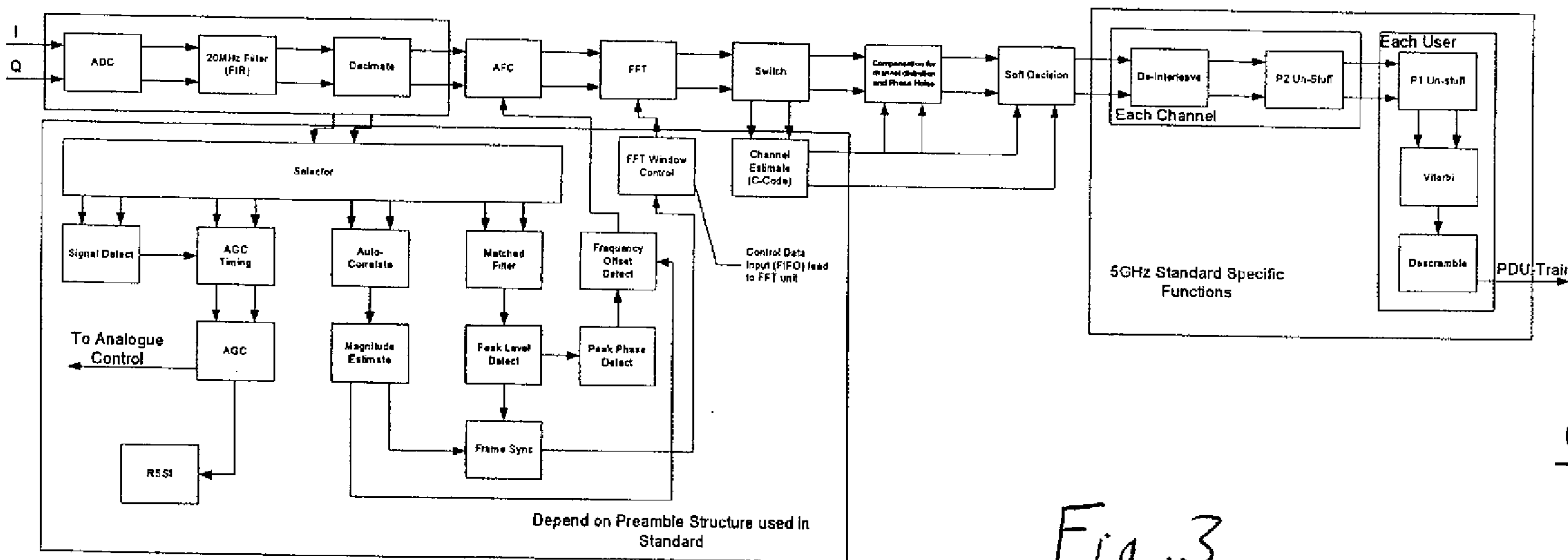
Fig. 1



Group 1
Group 2
Group 3
Group 4
203
202
201
200
PREAMBLE
Time domain (OFDM symbol number)
1
5
10
15
20
25
Ch. -26~-22 data
Ch. -21 pilot
Ch. -20~-8 data
Ch. -7 pilot
Ch. -6~-1 data
Ch. 0 DC
Ch. 1~6 data
Ch. 7 pilot
Ch. 7~20 data
Ch. 21 pilot
Ch. 22~26 data
Frequency domain (OFDM carrier index)

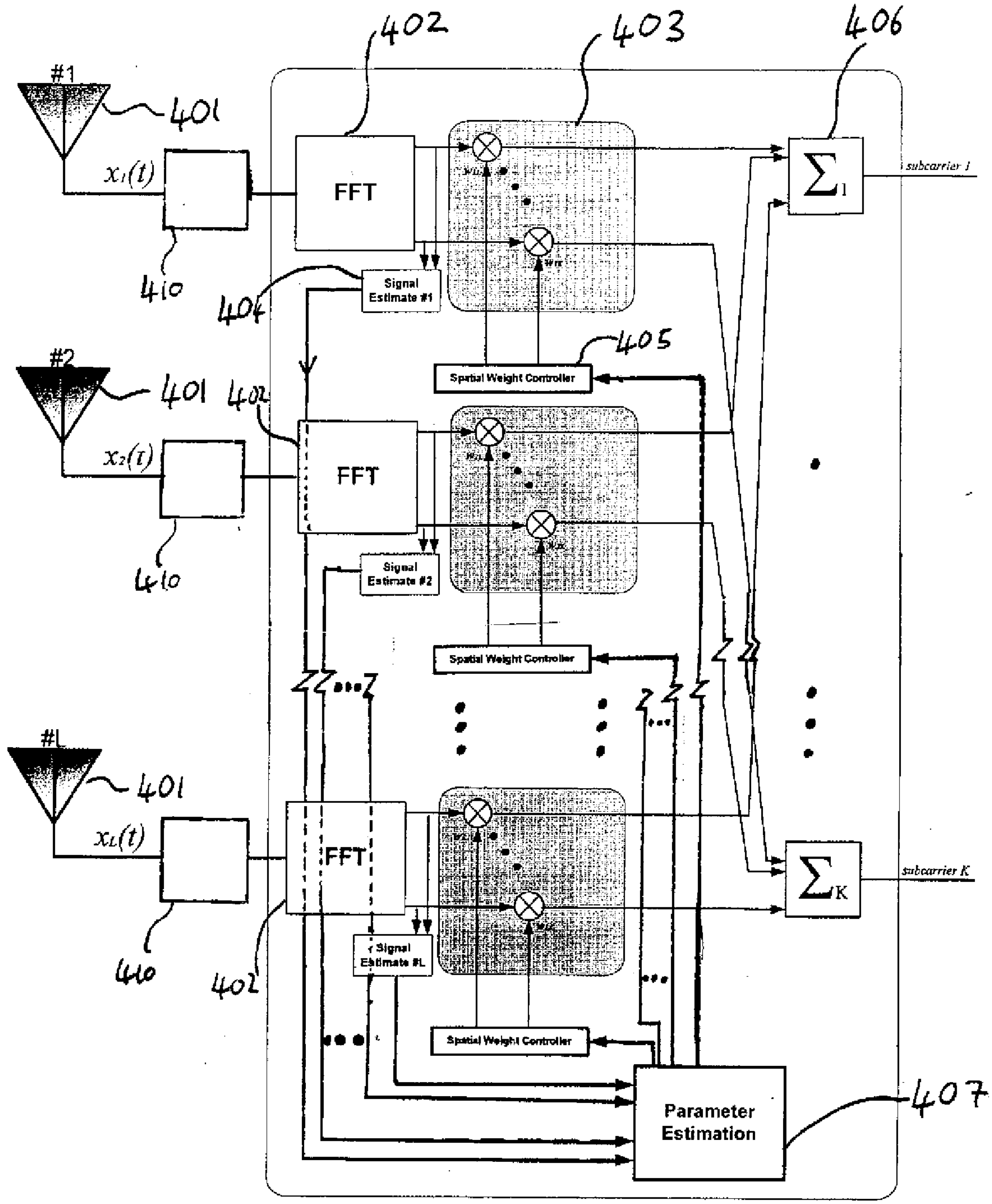
Fig. 2

3/6



I
Q
ADC
20MHz Filter (FIR)
Decimate
AFC
FFT
Switch
Soft Decision
De-interleave
P2 Un-Stuff
Each Channel
Each User
P1 Un-stuff
Viterbi
Descramble
PDU-Train
5GHz Standard Specific Functions
Selector
Signal Detect
AGC Timing
AGC
To Analogue Control
RSSI
Auto-Correlate
Magnitude Estimate
Matched Filter
Peak Level Detect
Peak Phase Detect
Frame Sync
Frequency Offset Detect
FFT Window Control
Channel Estimate (C-Code)
Control Data Input (FIFO) feed to FFT unit
Depend on Preamble Structure used in Standard

Fig. 3



402
403
406
#1
401
$x_1(t)$
FFT
subcarrier 1
410
404
Signal Estimate #1
405
Spatial Weight Controller
#2
401
402
$x_2(t)$
FFT
410
Signal Estimate #2
Spatial Weight Controller
#L
401
$x_L(t)$
FFT
subcarrier K
410
402
Signal Estimate #L
Spatial Weight Controller
Parameter Estimation
407

Fig. 4

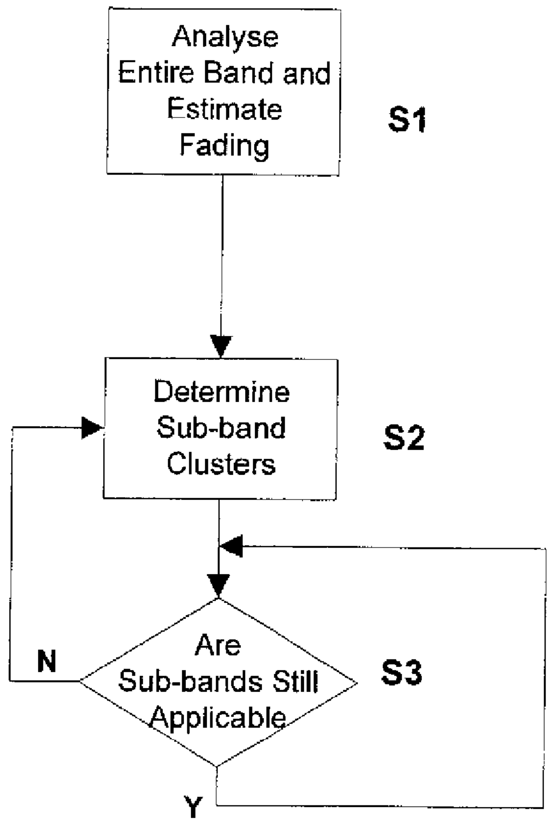

Analyse Entire Band and Estimate Fading
S1
Determine Sub-band Clusters
S2
Are Sub-bands Still Applicable
S3
N
Y

Fig. 5

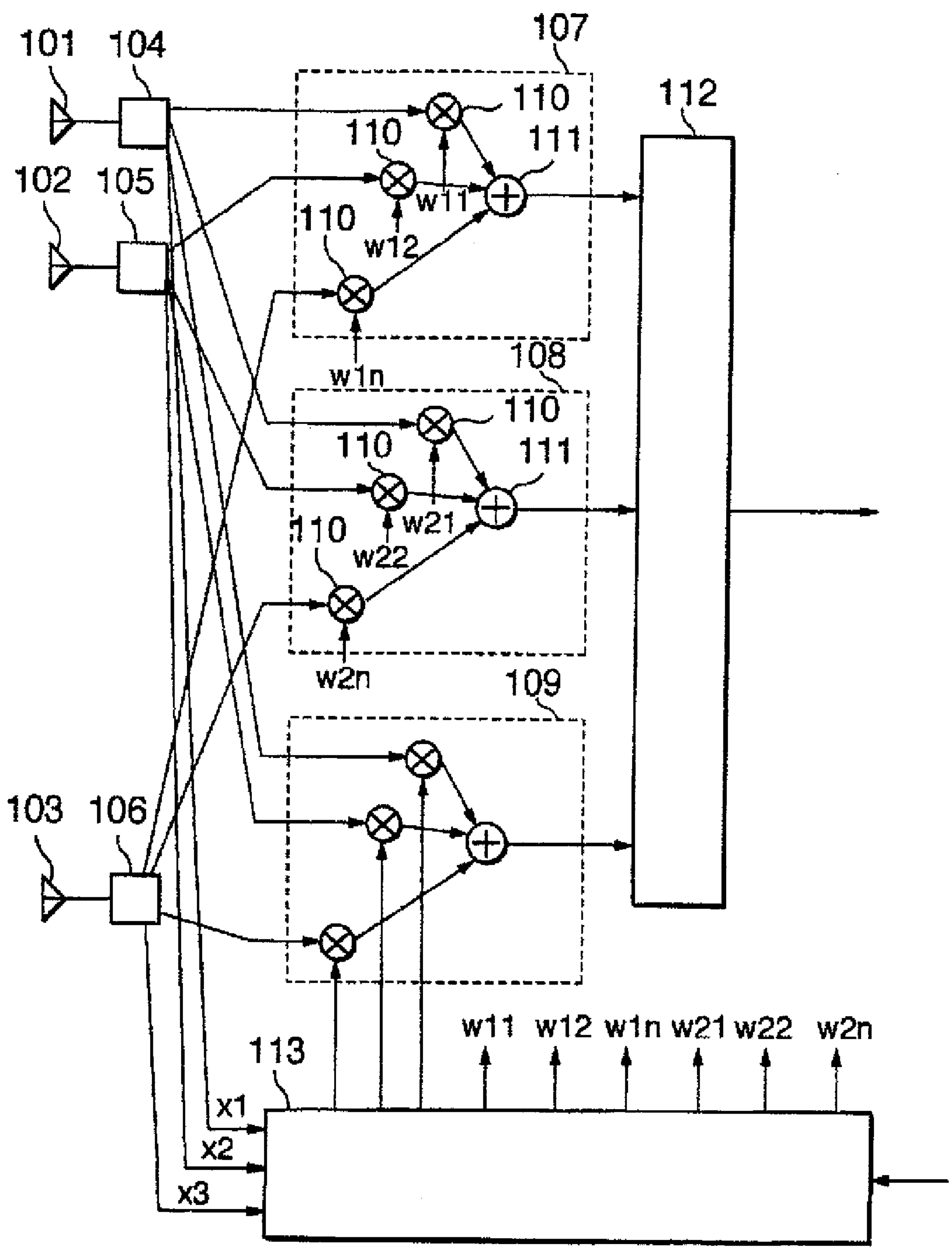

101
104
102
105
103
106
107
108
109
110
111
112
113
w11
w12
w1n
w21
w22
w2n
x1
x2
x3

Fig. 6

M&C Folio: GBP83498

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ADAPTIVE ANTENNA

The present invention relates to a multi-carrier or Orthogonal Frequency Division Multiplexing (OFDM) system employing multiple antennas. In particular, the present invention relates to the dynamic clustering of sub-carriers in the receiver.

There are now a number of systems which operate at high data rates such as multi-media applications like Wireless Local Area Network (WLAN); Wireless Personal Area Network (WPAN), e.g. Bluetooth; etc. Various arrangements have been proposed in order to facilitate such high data rates in a reliable and practical way. However, there are a number of problems associated with high data rate wireless channels, particularly due to multipath. This is especially problematic indoors where the environment is particularly dispersive due to the large number of objects and surfaces as well as the dynamic nature of structures, particularly people, moving about. Consequently, it can become extremely difficult to extract and retrieve the original information reliably and without extremely complex processing. This results in extremely complex receivers which must be capable of estimating and compensating for the multiple versions of the original signal arriving at the receiver due to the variation in the path length of each version.

In order to counter this, a multi-carrier approach has been proposed in which the original data stream is separated into a series of parallel data streams, each of which is modulated and transmitted with a different frequency, generally within the same band. This allows the relative size of the transmitted symbols to the multipath delay to be much larger and so inter symbol interference is reduced. One particularly advantageous system, which utilises multiple carriers is Orthogonal Frequency Division Multiplexing (OFDM).

OFDM is very effective at overcoming the problems of fading and multipath. This is achieved by dividing a frequency selective fading channel (i.e. a channel where the fading characteristics at one frequency are likely to be different to those at neighbouring frequencies i.e. the profile of the received signal against frequency is not flat) into a number of flat fading sub-bands such that the profile within the sub-bands is approximately flat. These sub-bands relate to the OFDM sub-carrier frequencies.

Figure 1 shows an example of the layout of a transmitter 10 and receiver 20 for an OFDM system. In the multi-antenna receiving unit 21, each of the antennas receives a signal which is fed to an analogue to digital converter 22 and then into a serial to parallel converter 23 to separate the individual sub-channels. The sub channels are then processed through a Fast Fourier Transform (FFT) 24. Finally, the signals are converted from a plurality of parallel signals into a serial signal for each sub-channel and the coded data extracted.

In such an OFDM receiver system, it is possible to apply adaptive beamforming weights at various points in a receiver as shown in Figure 1. However, the effectiveness of these weightings will largely depend on the stability and coherency of the propagation channel. If the channel undergoes flat fading, then it can be regarded as being a narrowband channel and a single set of weights can be applied at radio frequency (RF) or intermediate frequency (IF), to the received signal just after the antenna array 21, i.e. at position (1) in Figure 1. Alternatively, the weights may be applied after the analogue to digital unit 22 at position (2) in Figure 1. Both of these positions should be sufficient for optimum spatial processing.

However, in wideband systems operating at high data rates such as WLAN, WPAN, etc. where bandwidths of 10 MHz or higher may be required and/or systems operating in highly dispersive environments, signals will occupy a spectrum in excess of the coherence bandwidth, i.e. there will be significant variation in the quality/signal strength of the channels across the bandwidth. Consequently, it is unlikely that a single set of weights (i.e. as in narrowband beamforming) would be satisfactory for beamforming.

One way to overcome this problem is to process the received data and apply weightings in the receiver for each sub-carrier, after the FFT 24, i.e. at position (3). However, this is very processor intensive. Figure 6 shows an example of a receiver. In this system, the signal is received by antennas 101,102,103. Pre-processing units 104,105,106, carry out downconversion, A to D conversion, serial to parallel conversion and FFT processing. The outputs are then fed into an array of adaptive signal processing devices 107,108,109 which include a plurality of multipliers 110 which multiply each of the received signals by a weighting value w determined by a weight determining unit 113. Each of the weighted signals from the multipliers is then summed 111 to provide an output signal. The output signals from each of the weighting units is then fed to a combining unit 112 which extracts a data signal in which the delayed signals and interference signals have been removed from the received signal.

However, in the example shown, the receiver has L antennas and the number of sub-channels that each antenna receives is N. Therefore, the total number of weighting units required is L x N. This can lead to a very large number of multipliers 110 being required. For example, in the HIPERLAN system, there are 48 data sub-carriers and 4 pilot sub-carriers (N=52); there is also a DC channel (CH0) which does not carry data. This means that the receiver is complicated and this in turn results in the receiver being expensive and potentially subject to reliability problems. In addition, the weighting is normally implemented in software an so processor demand is extremely high, again resulting in high expense or poor performance. If the processing to determine the weighting to be applied is unduly complex, then it may take a significant amount of time to complete. During this time, the channel parameters may have changed significantly and so the calculated weightings could be inappropriate. Under these circumstances, the weighting produced would always out of date and hence poor performance will result where the characteristics of the channel change rapidly with time.

One way to reduce the processor demand, is to divide the operating bandwidth into a number of sub-bands and then select one sub-carrier from within each sub-band on

which to base all calculations. This method relies upon each sub-band behaving generally as a narrowband, i.e. that the sub-band effectively undergoes flat fading. In other words the chosen sub-carrier is accurately representative of the fade within the sub-band as a whole. However, without prior knowledge of the operational environment, it is difficult to know to what extent the operating band should be divided up. Where the sub-bands are chosen to be large there is a danger that the chosen sub-carrier would not be sufficiently representative of the sub-band and performance would be degraded. In contrast, if the number of sub-bands is chosen to be large, whilst the representative sub-carrier is likely to be accurately representative of the sub-band, the amount of processing required is disadvantageously high.

EP-A2-0,852,407, which relates to current standards for 5 GHz WLANs, suggests reducing the total number of adaptive signal processing units and hence the number of weighting units to improve the receivers by reducing the complexity. The document describes dividing the operating band into four equal sub-bands each having a 'pilot' sub-carrier.

An example of this arrangement is shown in Figure 2 where the operating band is divided into fifty-three channels or sub-carriers (i.e. as in HIPERLAN), these are then divided up into four separate groups each defining a sub-band. Each sub-band includes a sub-carrier which acts as a pilot for the group. The pilot channels do not carry signal data but contain a predetermined sequence for use in equalising the received signal by comparing the received signal to an expected signal. Weighting for the received signals is determined using the pilot sub-carriers and is then applied to each sub-carrier in the respective sub-band. As indicated above this system relies upon flat fading over the sub-bands which in the case of the above referenced document are of the order of 5 Mhz in size.

If the bandwidth of the system is increased such that the sub-bands have considerably greater bandwidth, for example in the region of 10 Mhz, then the likelihood that the sub-

bands will have flat fading is considerably reduced particularly where the environment is such as to give strong multipath interference, e.g. indoors.

Therefore according to the present invention there is provided an adaptive weighting system comprising:

frequency domain transform means for converting a received signal to a plurality of sub-channels;

banding means for allocating a sub-channel to a sub-band based upon a determination of the coherency of some or all of said sub-channels;

weight calculation means for determining a weighting for each sub-band; and

weighting means for applying the respective determined weight for each sub-band to the or each sub-channel of the sub-band.

The present invention further provides a method of processing sub-channels of a received broadband signal comprising:

transforming a received signal into a frequency domain signal,

determining the coherency of each sub-channel of the signal relative to other sub-channels;

allocating each sub-channel to one or more sub-bands based upon said determination of the coherency;

determining weights to be applied to each sub-band based upon a determined coherency of the sub-band; and

applying the determined weights for a sub-band to each of the sub-channels in that sub-band.

The banding means preferably allocates a sub-channel to a sub-band based upon the coherency of that channel being within a predetermined amount of the other sub-channels which have already been allocated to that sub-band. The maximum difference in coherency is preferably within 3dB of the other sub-channels of the sub-band. The maximum difference in coherency is preferably within 0.5 degrees of the other sub-channels of the sub-band.

The system preferably monitors the coherency of the sub-channels within a sub-band to ensure that they continue to remain within a certain range of the other sub-channels of the sub-band. The range may be the same as the predetermined amount or it may be larger.

The present invention applies the same weighting to each member of a sub-band. This means that only one weighting value must be calculated for each sub-band, considerably reducing the processing demand compared to calculating a weighting value for each sub-channel. This in turn allows the weightings to be determined more quickly and so there is less delay between the signal being received and the appropriate weightings being determined. This ensures that the weightings are more up to date and hence accurate.

A specific embodiment of the present invention will now be described in detail by reference to the drawings, in which:-

Figure 1 shows a block diagram showing a typical arrangement of transmitter and receiver for use in an OFDM system;
Figure 2 shows a representation of a system in which the operating band is divided into sub-bands;
Figure 3 shows an overview of a typical OFDM receiver;
Figure 4 shows a schematic representation of a basic receiver architecture for broadband adaptive antenna weight calculation;
Figure 5 shows a flow diagram of the operation of the OFDM receiver of the present invention; and
Figure 6 is a schematic diagram showing the structure of a conventional adaptive antenna system.

Figure 3 shows the functional layout of the baseband section of an OFDM receiver used in the present invention. Many of the functions of the receiver shown as well known

and so only a brief explanation is given here. The received signal is quadrature (I and Q) downconverted, amplified and filtered (not shown) before being (over-)sampled by the A/D unit. The digital over-sampled signal is then filtered and decimated. The over-sampling of the signal at the start aids the digital filtering process, after which it is then rate reduced to the required/expected sample rate. It is assumed in this case that the system provides for a preamble of some sort in every burst within a frame (MAC frame). In the case of HIPERLAN, each frame comprises a preamble portion which is made up of three basic OFDM symbols denoted her as A,B and C. A and B (or even C) symbols can be observed (recovered) in the time domain (pre-FFT) and used to establish the frame and frequency synchronisation (as well as set the FFT window for the data that follows these symbols) through some correlation process. The automatic gain control (AGC) settings (not shown -- prior to ADC) can also be established. It is possible to pass the C symbol through as a complete symbol to the FFT. Knowing what this symbol is in advance (and assuming adequate synchronisation), the channel variation can be estimated on a sub-carrier basis post-FFT. The C symbol would be 'switched' out to estimate the channel compensation (rotation of the symbols in the sub-carriers). However, this same channel estimation could be used in the sub-carrier grouping procedure.

Alternatively, pilots can be selected post-FFT and used to estimate the channel over time. The pilots have known symbols in them and are processed to identify what symbol is received (I' and Q') and what symbol was expected (I and Q). If $I' \neq I$ and $Q' \neq Q$ then you can calculate the phase rotation and amplitude change required to make them equal. Given four estimates of these values (4 pilots) you can estimate or interpolate the amplitude and phase rotation required for all the intermediate sub-carriers. This is known as a one-tap equalisation which is a simple if not crude way to determine the correction needed for all the channels. The determined value of the sub-carrier's amplitude and rotation (I and Q) correction value can then be applied. The remainder of the system carries out the unpacking and unscrambling of the data to the relevant bits.

It should be noted that the preamble symbols are there for 'training' or synchronisation purposes i.e. they are known at the receiver so that it can form an estimate of the influences of the channel for an equaliser or even smart antenna weight calculation.

The basic operation of the receiver system will now be described with reference to figure 5. Data is transmitted in blocks of data known as symbols. These symbols typically comprise a guard interval for reducing inter-symbol interference as well as a useful data part. Each symbol is transmitted on a sub-channel using the sub-carriers referred to above. The data to be transmitted is divided up into symbols which are then sent on each of the sub-channels and re-constructed at the receiver. In addition, the series of symbols transmitted generally include one or more preamble symbols, as indicated above for providing control and synchronisation information etc.

Initially, at step S1, the receiver performs a first estimate of the variation of the fading across the entire band using the preamble symbol, in the frequency domain. An estimate of the band is obtained in quadrature (I and Q) information. The 'flatness' of the received power across the sub-carriers and phase differences can be obtained.

At step S2, the receiver divides the sub-carriers across the band into groups of sub-carriers to form sub-bands. The sub-carriers are grouped with other sub-carriers which are within a certain range of each other. The received power and phase differences do not have to be absolutely equal but just sufficiently close to be within a certain range. The range may be varied depending upon the circumstances. By making the range small, the groups of sub-carriers will be small but very coherent. In contrast, if the range is large, the groups can be much larger but they could be less coherent. The range is therefore selected according the available processing power, required reception quality and so on. A typical value for the range could be that if sub-carriers are within 3dB of received power and 0.5 degrees of variation.

Then for each group, the appropriate weightings are determined based on one sub-carrier of the group, and those weightings are applied for all of the sub-carriers in that

group. The weights are determined, as indicated above, such that the level of coherence within each sub-band is suitable for the efficient and effective application of a single set of adaptive array weights for all sub-carriers based on a calculation for a single sub-carrier's performance, i.e. all sub-carriers within the group can be considered to be within the coherence bandwidth of the current channel.

As indicated above, the appropriate weightings are determined based on one sub-carrier of the group. The chosen sub-carrier should in principle be representative of the group so that when the weightings are applied, they are applicable to all members of the group. In practice, it would be very processor intensive to test each sub-carrier to determine which one was most representative. Therefore, the selection of the sub-carrier is selected using other criteria. A number of criteria can be used. If the sub-carrier includes a pilot carrier then this can be used as this has a known data content and will provide a fairly accurate representation of the fading of that channel and hence the group. Clearly not all channels will include a pilot. If you were working with an adaptive antenna algorithm that worked on signal characteristics (rather than data content) then it would probably be satisfactory to pick whichever sub-carrier was central to the group. In essence, unless there is significant fading across the group, then the choice of sub-carrier is not critical and by their very nature, the groups are selected not to have significant fading across them.

By using the same weightings for more than one sub-carrier, the total number of calculations needed to determine the weights for each of the sub-carriers can be reduced. Once the weightings have been determined, the weights can be applied to the sub-band for as long as the sub-band retains sufficient coherency. The coherency of the band will vary over time, particularly due to environmental variations. In order to ensure this, the coherency the sub-carriers is periodically monitored. The frequency of checking the coherency will depend on the variability of the environment. If the environment is generally stable for long periods then checking can be less frequent but in a highly variable environment, regular checking will be required.

Where the data transfer traffic is very bursty, i.e. data is received in short bursts with possibly long periods of no data in-between then it may not be possible to monitor constantly and it may only be possible to form a single estimate of the groups. In contrast where communication is fairly constant, such as in voice communication, it may be possible to make several estimates for determining the groups.

Furthermore, with bursty traffic, if an estimate is made, this may out of date by the time the next estimate is made in a subsequent burst if the period between bursts is longer than the period in which a channel remains coherent. Signal interference will vary over time due to changes in environmental influences particularly if the transmitter or receiver is moving. Consequently where the data is bursty, there may be insufficient time to make more than one estimation. Thus the estimation may have to be carried out on the basis of a few or even a single frame of data. However, in the case of adaptive algorithms, i.e. ones that take a period of time (several samples rather than one) to converge to an answer, if the input conditions change, then clearly they need to re-converge to take account of the new environment. However, this may potentially be from a known converged point which may increase the convergeance rate. In other words, the weights used for a particular transmitter could be used as a starting point for the next set of measurements and determination of sub-banding groups. So rather than starting from scratch each time, the process can begin based upon the conditions determined in the previous determination. Hence they can adapt or are adaptive. These algorithms usually require some sort of training sequence (which could be a preamble) and so work in a similar way to equalisers, even using the same or similar algorithmic techniques.

The coherency of the groups is periodically monitored (step S3) to ensure that they remain within predetermined limits. The coherency may be checked to ensure that the members of the group are still within the range of values used to determine the groups initially. Alternatively the range may be slightly larger to allow some decrease in the coherency (to avoid a very small reduction in coherency precipitating a complete re-assessment of the groupings). If the coherency of any of the sub-carriers in a group is

below the predetermined level, then all the groups are re-assessed as above and the sub-carriers reallocated to a new set of groups. Alternatively, just the weights could be re-assessed if the coherency is still within another limit. As a further alternative, rather than reassess all the groups, if only one group is out of range then it may be possible to divide just that group into two or more new groups which in themselves are sufficiently coherent. Although this increases the number of groups, this may be acceptable as an interim measure to avoid having to re-assess all of the groups.

As indicated above, the determination of the sub-banding is based upon calculations on the preamble channels. However, the calculations can also be on the pilot channels provided in the received signal or on known modulation or signal characteristics such as direction of arrival estimation (Fourier method, ESPRIT, MUSIC) and blind optimal combining (least squares, recursive least squares, sample matrix inversion etc.)

An explanation of the operation of a circuit according to the present invention will now be described with reference to figure 4. The arrangement shown is similar to the prior art systems in that it comprises a plurality of antennas 401 for receiving a transmitted signal. A receiver unit 410 carries out pre-processing of the signal from each antenna to filter the received signal and carry out downconversion. The signal is also prepared for processing in the receiver unit 410 by carrying out A to D conversion. The received signals from each receiver unit 410 are then transformed by a respective FFT 402. The output from each FFT is then passed to an adaptive weighting unit 403. The output from each FFT is also passed to a signal estimating unit 404 associated with each FFT. The signal estimating unit 404 is primarily to carry out channel estimation for the allocation of each channel to a group. This can comprise a correlator for part of the preamble that has been passed through the FFT or the pilot symbols for a longer estimation method (the pilot symbols run throughout the burst whereas the preamble only occurs for the first few symbols of a burst, see figure 2). In figure 2, the bottom axis is frequency and the side axis represents time in symbol durations. The preamble lies at the start of the transmission, and the pilots, which are shown shaded, run through the entire burst. The remainder are the normal sub-carriers for carrying payload data.

The parameter estimation units 407 receive the results of the channel estimation from the signal estimators 404. Based on the determined groups, this unit calculates the weightings for each group according to the selected representative sub-carrier from each group. The parameter estimation circuit 407 controls the spatial weight controllers 405, associated with the adaptive weighting unit 403 in each branch, to output the appropriate weightings for each channel of the adaptive weighting unit. The adaptive weighting units apply the determined weightings to each of the channels of data received from the FFTs. The weighted outputs are then provided to respective adders 406. An adder 406 is provided for each sub-carrier rather than each branch. The adders 406 sum the appropriately weighted channels from each FFT. The output from each adder represents a filtered sub-carrier signal derived from the plurality of delayed signals received by the antennas.

Although, the units described above are indicated as separate. In fact most of the calculations will be carried in a processor such as a DSP and not in discrete hardware units although this is not essential. It is typical for all operations after the A to D conversion to be carried in a software controlled DSP.

The present invention is particularly advantageous when used with more complex algorithms, particularly ones where forming a channel estimate that will require an amount of processing time dependant on the number of sub-carriers. For example, if you are using an 'adaptive' scheme that is going to need hundreds of samples to converge, it is clearly advantageous to limit the number of sub-carriers it needs. In this way, reducing the number of sub-carriers by grouping, will advantageously reduce the processing time to determine the weightings. In contrast, if a 'simple' algorithm is used then the grouping process may actually add extra complexity to the whole procedure.

These systems are intended to cover operation over relatively large overall bandwidths. For example, HIPERLAN may operate in the bands 5.15-5.35GHz AND 5.47 to 5.725GHz, so with both these bands it may arise that the channel is extended. However, one set of sub-carriers may be a long way off the next set, in frequency. Consequently,

when the groups are being determined it may be necessary to apply some bounding conditions. This avoids the system trying to group together spectrally disparate sub-carriers.

CLAIMS:

1. An adaptive weighting system comprising:

frequency domain transform means for converting a received signal to a plurality of sub-channels;

banding means for allocating a sub-channel to a sub-band based upon a determination of the coherency of a plurality of said sub-channels;

weight calculation means for determining a weighting for each sub-band; and

weighting means for applying the respective determined weight for each sub-band to the or each sub-channel of the sub-band.

2. An adaptive weighting system according to claim 1 wherein the banding means determines the coherency of each sub-channel.

3. An adaptive weighting system according to claim1 or 2 further comprising control means for controlling the banding means to check the coherency of the sub-channels of the sub-bands to determine that the maximum difference in the coherency of the group is below a predetermined amount.

4. An adaptive weighting system according to claim 3, wherein the control means is adapted to re-assign the sub-channels to new sub-bands if it is determined that the maximum difference in the coherency of the group is below a predetermined amount.

5. An adaptive weighting system according to any one of claims 2 to 4 wherein the coherency measurement is based upon received power and/or phase difference.

6. An adaptive weighting system according to any one of claims 2 to 5 wherein the predetermined amount is 3dB for the received power and/or 0.5 degrees for the phase difference.

7. An adaptive weighting system according to any one of the preceding claims wherein the banding means allocates a sub-channel to a sub-band such that the

coherency of other sub-channels in the sub-band is within a predetermined amount of the coherency of the allocated sub-channel.

8. An adaptive weighting system according to any one of the preceding claims wherein the weight calculation means determines the weighting according to a determined coherency value for the sub-band.

9. An adaptive weighting system according to any one of the preceding claims wherein the weight calculation means determines the weighting according to the coherency of a selected one of the sub-channels of a sub-band.

10. A method of processing sub-channels of a received broadband signal comprising:

transforming a received signal into a frequency domain signal,

determining the coherency of each sub-channel of the signal relative to other sub-channels;

allocating each sub-channel to one or more sub-bands based upon said determination of the coherency;

determining weights to be applied to each sub-band based upon a determined coherency of the sub-band; and

applying the determined weights for a sub-band to each of the sub-channels in that sub-band.

11. A method of processing sub-channels of a received broadband signal according to claim 10 wherein the determination of the coherency of a sub-band is based upon the coherency of a selected sub-channel of that sub-band.

12. A method of processing sub-channels of a received broadband signal according to claim 12, wherein determining the coherency of a sub-channel comprises determining the received power and/or phase difference of the sub-channel.

13. A method of processing sub-channels of a received broadband signal according to claim 10,11 or 12, wherein said determination of said coherency comprises determining that the coherency of the sub-channel is within a predetermined amount of the coherency of the other sub-channels in the group.

14. A method of processing sub-channels of a received broadband signal according to claim 13, wherein the predetermined amount is 3dB for the received power and/or 0.5 degrees for the phase difference.

15. A processor for processing sub-channels of a received broadband signal, the processor being adapted to:

transform received signals into frequency domain signals,

determine the coherency of each sub-channel relative to other sub-channels;

allocate each sub-channel to one or more sub-bands based upon said determination of the coherency;

determine weights to be applied to each sub-band based upon a sub-channel within the sub-band; and

apply the determined weights for each sub-band to each of the sub-channels in the sub-band.

16. An adaptive weighting system substantially as described herein with reference to the accompanying drawings.

17. A method of processing a received broadband signal substantially as described herein with reference to the accompanying drawings.

18. A processor substantially as described herein with reference to the accompanying drawings.

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Claims searched: 1-18

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Patents Act 1977
Search Report under Section 17

Databases searched:

UK Patent Office collections, including GB, EP, WO & US patent specifications, in:

UK Cl (Ed.S): H4P (PAN, PAR, PAX, PRE)

Int Cl (Ed.7): H04L: 27/26

Other: Online: EPODOC, JAPIO, WPI

Documents considered to be relevant:

Category	Identity of document and relevant passage		Relevant to claims
A	GB 2356769 A	[SECRETARY OF STATE FOR DEFENCE]	
X	EP 1037303 A1	[MOTOROLA] See Figure 1 and paragraphs 22,40,45-47.	1-15
A	EP 0851642 A2	[SHARP]	
A	WO 97/40608 A1	[AMATI]	
A	US 6249250 B1	[NAMEKATA]	

X	Document indicating lack of novelty or inventive step	A	Document indicating technological background and/or state of the art.
Y	Document indicating lack of inventive step if combined with one or more other documents of same category.	P	Document published on or after the declared priority date but before the filing date of this invention.
		E	Patent document published on or after, but with priority date earlier than, the filing date of this application.
&	Member of the same patent family		

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(54) **COMMUNICATION EQUIPMENT**

(57)Abstract:
PROBLEM TO BE SOLVED: To make reception data in a base station hold the orthogonal relation of an orthogonal code and to synchronize reception signals by compression- encoding the advance/delay information of the time calculated so as to match the synchronization of the orthogonal code of the reception signals from respective mobile stations and transmitting it to the mobile stations.
SOLUTION: Output is passed through a parallel/serial converter 407 and turned into serial data. To the data, noise equivalent to a time difference with the reception data from the surrounding mobile station is superimposed. The data are judged by using a judgement device 409. Noise signals are used and delay time is predicted by using a delay time control signal predicting device 411. The predicted delay time difference is encoded and transmitted from the transmitter of the base station to the respective mobile stations. The signals are turned into input signals to a receiver and the timing of the transmission signals of the respective mobile stations is adjusted. Thus, the orthogonal codes used as the spreading signals of the data in the respective mobile stations are synchronized.

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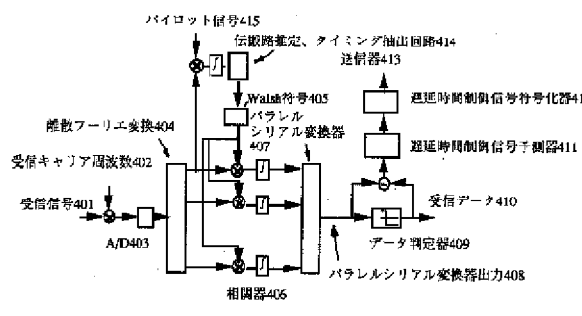
最終頁に続く

(54)【発明の名称】　通信装置

(57)【要約】

【課題】移動通信システムのセル内での移動局間の干渉が小さくなり、セル内の加入者を増大でき、周波数効率を増大できる装置を提供すること。

【解決手段】入力データを一定の符号によって拡散する機能と、拡散された信号を一つ以上の周波数チャネルを用いて変調して送信する機能を有する送信器と、各々の周波数チャネル信号に対して復調する機能と送信器と同等の符号を用いて相関検波を行う機能を有する受信器で構成された通信装置において、受信器が、同時に一つ以上の送信器からの信号を受信した受信時刻を決定する機能と、受信時刻が一定になるように各送信器の送信時刻を決定する機能と、各送信器からの送信時刻を各送信器に対して送信する機能を有し、送信器は、受信器からの送信された送信時刻に従って、送信データを送信する機能を有することを特徴とする通信装置。



【特許請求の範囲】

【請求項１】入力データを一定の符号によって拡散する機能と、拡散された信号を一つ以上の周波数チャネルを用いて変調して送信する機能を有する送信器と、各々の周波数チャネル信号に対して復調する機能と、送信器と同等の符号を用いて相関検波を行う機能を有する受信器で構成された通信装置において、受信器が、同時に一つ以上の送信器からの信号を受信した受信時刻を決定する機能と、受信時刻が一定になるように各送信器の送信時刻を決定する機能と、各送信器からの送信時刻を各送信器に対して送信する機能を有し、送信器は、受信器からの送信された送信時刻に従って、送信データを送信する機能を有することを特徴とする通信装置。

【発明の詳細な説明】

【０００１】

【発明の属する技術分野】移動通信方式、特に通信装置の構成に関するものである。

【０００２】

【従来の技術】従来、この種の方式は「Ｔ．Ｍｕｌｌｅｒ 他“Ｃｏｍｐａｒｉｓｏｎ ｏｆｄｉｆｆｅｒｅｎｔ ＤｅｔｅｃｔｉｏｎＡｌｇｏｒｉｔｈｍｓ ｆｏｒ ＯＦＤＭ－ＣＤＭＡ ｉｎ Ｂｒｏａｄｂａｎｄ Ｒａｙｌｅｉｇｈ ｆａｄｉｎｇ”のＩＥＥＥＶＴＣ９５」の第８３５頁第１図に開示されたものがある。これは入力データをＷａｌｓｈマトリクスで変換し、ＩＦＦＴで変換して出力する構成であった。変調器の構成を第１図に示す。復調器の構成を第２図に示す。

【０００３】

【発明が解決しようとする課題】しかしながら、従来の構成では、移動通信システムの基地局の送信器から、システム内の複数の移動局に対して、入力データを直交符号で拡散して送信し、同時に送信し、各移動局は各データの直交符号により直交関係を保持したまま同期して受信することは可能であるが、逆に各移動局から送信するデータを基地局で受信する場合、基地局と各移動局との距離は各々異なることから、入力データを直交符号で拡散して送信しても、基地局での受信データは直交符号の直交関係を保持して受信信号の同期を取ることができない。

【０００４】

【課題が解決するためにの手段】各移動局から直交符号で拡散して送信されるデータの受信時刻を、各移動局毎に基地局で測定する機能と、基準時刻からのズレ量を移動局毎に決定する機能と、基地局において各移動局からの受信信号の直交符号の同期が合うように計算された時刻の進み遅れ情報を、圧縮符号化して移動局へ送信する機能と、各移動局は直交符号で拡散して送信するデータの時刻を、受信した時刻の進み遅れ情報に合わせる用に調整して送信する機能を有することを特徴とする通信装置。

【０００５】

【発明の実施の形態】最良と考える本発明の実施の形態（発明をどのように実施するか）を、図面に基づいてその作用効果を示して簡単に説明する。

【０００６】本発明の移動局側の送信器の構成において、入力データを直並列変換器を通して複数の並列データに分解し、各々を固有の直交符号を用いて拡散し、拡散したデータを直交関係にあるキャリア周波数で変調して送信する。従って、１つのチャネルのデータ速度は、チャネルの数分だけ低くなる。

【０００７】直交符号はＷａｌｓｈ符号等の符号が用いられる。直交関係にあるキャリア周波数による変調は、複数の直交符号で拡散した信号を逆離散フーリエ変換し、その出力を並直列変換することで得ることができる。

【０００８】基地局の受信器の構成において、複数の移動局から送信された固有の直交符号で拡散されたデータを、同時に受信する。各移動局の信号の受信タイミングに従って、各々の信号を離散フーリエ変換し、各移動局固有の直交符号で相関を取る。相関出力をパラレルシリアル変換してデータ判定を行う。この判定データを用いて遅延時間信号予測器により、遅延時間を算出する。遅延時間は全部の移動局からの受信信号の受信タイミングのほぼ平均タイミングからのズレ時間となる。

【０００９】遅延時間をディジタル符号化して、基地局の送信器から移動局の受信器に送信する。移動局は、遅延時間を受信し、その時間に対応した時間に合わせながら送信タイミングを決定し、送信信号を送信する。

【実施例】

【００１０】本発明の具体的な実施例について図面に基づいて説明する。

【００１１】第一の実施例

１．構成の説明

図３はこの発明の第一実施例を示す、移動局側の送信器構成図であって、入力データ３０１はシリアルパラレル変換器３０２で、１対Ｎのパラレルデータに変換される。シリアルパラレル変換器３０２によって変換された各チャネルのデータの速度は入力データ速度の１／Ｎとなる。

【００１２】変換されたデータは拡散器３０４により、直交符号であるＷａｌｓｈ符号３０３を用いて拡散信号となる。チャネルの拡散信号にパイロット信号３１４を相加する。パイロット信号は複数のチャネルに相加してもよい。

【００１３】拡散信号は離散逆フーリエ変換３０５を通して出力される。離散フーリエ変換出力は、切り替え器３０６により各チャネルを連続データに変換し出力する。出力するタイミングは受信器３０７より受信したデータに従って行う。送信タイミングに従って出力した信号はＤ／Ａ変換器３０９を介して出力する。出力信号は

キャリア周波数により変調され送信する。

【００１４】図４は本発明の第一実施例を示す、基地局側の受信器の構成であって、受信信号４０１は、受信キャリア周波数４０２によって復調される。復調信号はＡ／Ｄ変換器４０３でディジタル信号に変換され、サンプルホールドされた後、離散フーリエ変換４０４によってＮチャネルのパラレル信号に変換する。図３の送信側において、パイロット信号が相加されたチャネル信号に対して、パイロット信号４１５と相関演算を行う。相関演算出力を、伝搬路推定、タイミング抽出回路４１４へ入力する。

【００１５】一方、各チャネル信号は、送信側と同じ直交符号４０５によって相関をとる。出力はパラレルシリアル変換器４０７を通してシリアルデータとなる。このデータには、回りの移動局からの受信データとの時間差に相当する雑音が重畳している。このデータを判定器４０９を用いてデータ判定する。雑音信号を用いて、遅延時間制御信号予測器４１１を用いて遅延時間を予測する。予測された遅延時間差を符号化して、基地局の送信器から各移動局に対して送信する。

【００１６】この信号は先の図３に示した、受信器３０７への入力信号となり、各移動局の送信信号のタイミングを調整する。

【００１７】２．動作の説明

図３の変調器の構成におけるタイムチャートを図６に示す。変調器は移動局側にある。一つの移動局の変調器について説明する。入力データ（ａ）のデータ速度をｄとする。データはＮチャネルのシリアルパラレル変換器によってパラレルデータ（ｂ）に変換される。各チャネル当たりのデータ速度はＮ分の１のｄ／Ｎとなる。

【００１８】各チャネルのデータを、各ユーザ毎に割り当てられた直交符号の一種であるＷａｌｓｈ符号（ｃ）によってＭ倍の拡散信号（ｄ）とする。ＭとＮは同じでもよい。図６では簡単のために、ＮとＭが等しい場合について示した。従って、拡散された後のチャネル信号の速度はｄとなる。ここで、一つの移動局に割り当てられるＷａｌｓｈ符号は一種類である。

【００１９】その内の、少なくとも一つのチャネル信号にパイロット信号（ｅ）を相加する。どのチャネルにパイロットを加えるかは予め定めておく。パイロット信号は、各ユーザ毎に割り当てられた直交符号と直交関係にあるＷａｌｓｈ符号を選択する。ここでは，パイロット信号との相加結果は示していない。

【００２０】拡散された各チャネル信号を離散逆フーリエ変換（ｆ）する。ここでは、チャネル１とチャネル２の実部と虚部の結果を示した。チャネル３，チャネル４の結果は省略した。

【００２１】離散フーリエ変換した出力（ｇ）は、切り替え器を用いて連続データとする。切り替え器の速度は、この例ではｄ／Ｎとなる。切り替え器の出力を送信するタイミングは、受信器から得た遅延時間制御信号の送信タイミング信号に従って決められる。遅延時間制御により送信タイミングを調整し、Ｄ／Ａ変換器を通して、送信キャリア周波数で変調して送信する。

【００２２】図４の復調器の構成におけるタイムチャートを、図６の結果と同様になる。復調器は基地局に設置され、複数の移動局からの信号を同時に受信する。各移動局からの受信タイミングは、基地局と移動局との相対距離によって異なる。

【００２３】一つの移動局からの受信信号について、動作説明を行う。受信信号はミキサーにより復調（ｇ）される。復調信号は離散フーリエ変換により各チャネル信号に変換（ｄ）される。変調器においてパイロットチャネルを相加されたチャネルにパイロット信号（ｅ）との相関演算を行い、無線回線の伝搬路推定を行うと同時にパイロット信号の受信タイミングを決定する。

【００２４】各チャネル信号は変調器で定められたＷａｌｓｈ符号（ｃ）を用いて、相関演算を行う。相関演算の相関タイミングはパイロット信号との相関演算で求められた受信タイミングを用いる。各チャネルデータ（ｂ）をパラレルシリアル変換してシリアルデータ（ａ）とする。シリアルデータからデータ判定を行い、受信データを得る。

【００２５】次に、遅延時間制御部について述べる。復調器は各移動局から同時に信号を受信するため、各移動局と基地局との相対距離により、受信タイミングにズレが生ずる。このズレが干渉雑音として、図４パラレルシリアル変換器の出力に相加される。

【００２６】図４では、この雑音量を評価する方法の一例として、パラレルシリアル変換器の出力とデータ判定出力との差から求めている。差の信号出力を次式で示す。

【数１】

$$e_k(t)=\hat{u}_k(t)-R_k(t)$$

ここで、$\hat{u}_k(t)$は判定データを表わす。

【００２７】パラレルシリアル変換器の出力をR_k（ｔ）で表わす。この信号を遅延時間制御信号予測器４１１へ入力する。遅延時間制御信号予測器では誤差信号を２乗平均を求める。

【数２】

$$J=E[e_k(t)^2]$$

遅延時間$T_{dk}^{(v+1)}$は次式を用いて更新する。

【数３】

$$T_{d,k}^{(v+1)}=T_{d,k}^{(v)}-\Delta T_d sign\frac{\partial J}{\partial T_{d,k}^{(v)}}$$

ΔT_dは遅延時間の更新量を表わす。νは今の遅延時間を表わす添字を表わす。ν＋１は次の時間での遅延時間の更新量を決める添字を表わす。

【００２８】実際には遅延時間をそのまま送った場合は

情報量が大きくなることより、遅延時間制御信号符号化器において、遅延時間を一定量進める、遅らせる、なにもしない、と情報量を圧縮して各移動局へ送信する。この符号化はシステムの構成により自由に選ぶことが出来る。

【００２９】図3に示した、各移動局は、この情報を受信器で復調して、予め定めた遅延時間幅３０８に基づいて決定し、各移動局からの送信タイミングを調整してデータを送信する。

【００３０】今までの説明では、遅延時間制御方法として、山上り法に基づいて説明したが、最小自乗法による等化器を用いても同様の動作することができる。

【００３１】今までの説明では、各チャネルが干渉を自動的に最小になるように、制御しているが、全てのチャネルに対して、共通の基準時刻を設定して、各チャネルの図４の遅延時間制御予測器４１１に入力し、その時刻に合うように遅延時間制御信号符号化器４１２が遅延時間信号を求めることにより制御することも可能である。

【００３２】今までの説明では、各移動局の最初の送信タイミングについて述べていなかった。移動局は基地局から送られてくるデータを、受信した時のタイミングでデータを送信してから、先に述べた遅延時間制御を行うことも可能である。

【００３３】また、先に述べた、自動的に干渉信号が最小となる方式と組み合わせることにより、各移動局と基地局の相対距離は大幅にずれていても遅延時間の制御は可能となる。

【００３４】また、干渉除去のためにガードインターバルを用いても良い。これは、本発明における必須要件ではない。

【００３５】今までの説明では、データをスペクトル拡散するマルチキャリア方式についての遅延時間制御を用いた通信装置について述べてきたが、スペクトル拡散装置そのもの対しても本方式を適用することができる。

【００３６】３．効果の説明

この第一実施例のように構成すると、移動局から基地局に対してデータを送信する場合、各移動局と基地局との相対距離に関係なく、各移動局でデータの拡散信号として使用した直交符号の同期をとることができる。これにより、誤り特性の優れた、データを拡散するマルチキャリア通信装置を構成できる。

【００３７】利用形態の説明

本実施例では無線回線で使用する装置とした例で説明したが、無線変調信号の代わりに光りでベースバンド信号を変調した光りファイバを用いた光通信装置にも適用可能である。また、無線変調信号の代わりに有線回線に用いるモデムアナログ変調を用いた有線通信装置にも適用可能である。

【００３８】キャリア周波数で変調を行ったが、振動子を用いることで水中通信装置への適用も可能である。

【００３９】

【発明の効果】本発明は上述のように構成したから、基地局において受信した、各移動局からの送信信号の干渉を減少することができ、耐雑音特性の優れた通信装置の実現ができる。

【図面の簡単な説明】

【図１】従来例のマルチキャリア変調器の構成を示す説明図である。

【図２】従来例のマルチキャリア復調器の構成を示す説明図である。

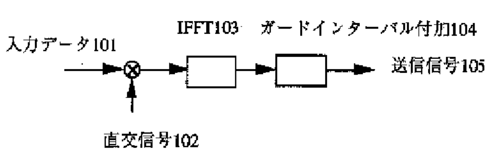
【図３】第一実施例の移動局側のマルチキャリア変調器の構成を示す説明図である。

【図４】第一実施例の基地局側のマルチキャリア復調器の構成を示す説明図である。

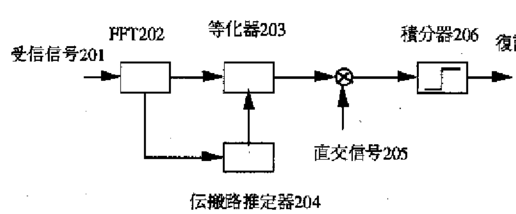
【図５】第二実施例の基地局側のマルチキャリア復調器の構成を示す説明図である。

【図６】第一実施例の送信器のタイムチャートである。

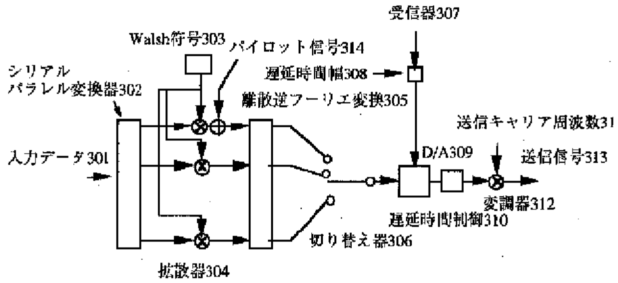
【図１】



【図２】

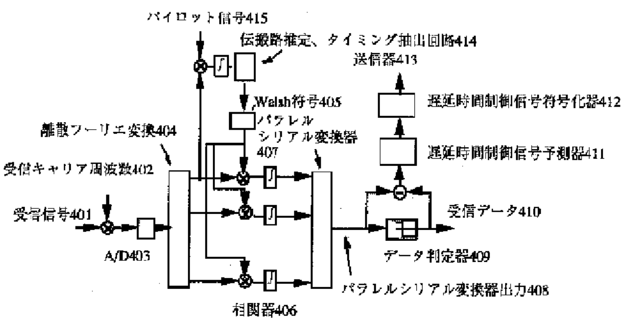


【図３】



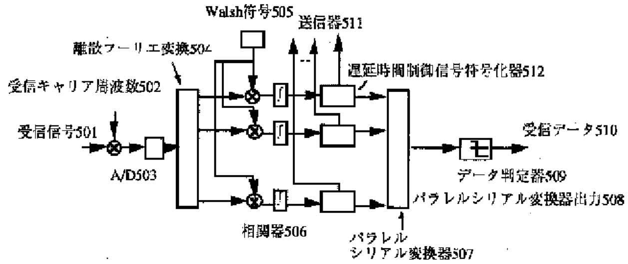
受信器307
Walsh符号303
パイロット信号314
遅延時間幅308
シリアル
パラレル変換器302
離散逆フーリエ変換305
送信キャリア周波数311
入力データ301
D/A309
送信信号313
変調器312
遅延時間制御310
切り替え器306
拡散器304

【図４】



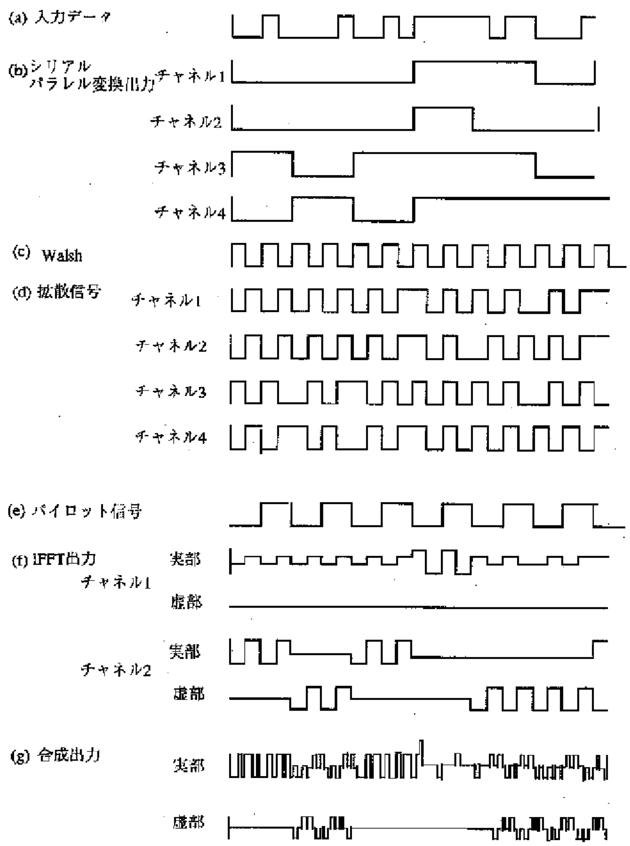
パイロット信号415
伝搬路推定、タイミング抽出回路414
送信器413
Walsh符号405
遅延時間制御信号符号化器412
パラレル
シリアル変換器
407
離散フーリエ変換404
遅延時間制御信号予測器411
受信キャリア周波数402
受信信号401
受信データ410
A/D403
データ判定器409
パラレルシリアル変換器出力408
相関器406

【図５】



Walsh符号505
送信器511
離散フーリエ変換504
遅延時間制御信号符号化器512
受信キャリア周波数502
受信信号501
受信データ510
A/D503
データ判定器509
パラレルシリアル変換器出力508
相関器506
パラレル
シリアル変換器507

【図6】



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JP9266466A DIGITAL TRANSMISSION SYSTEM

Bibliography

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Multi carrier digital transmission system for CATV network has sub-channel modulator provided to each terminal station to modulate symbol row to carrier signal and output baseband time series signal

Original Title

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Abstract

PROBLEM TO BE SOLVED: To provide the digital transmission system which can easily cope with an increase in terminal stations and is tolerant of monotone noise while effectively using a frequency band.

SOLUTION: At a terminal station 10, a plurality of subchannel carrier signals characteristic of the terminal are modulated by an SP converter 11, an inverse FFT 12, a PS converter 13, and a DA converter 14 with a symbol sequence to be sent to a center station 30 and a carrier signal is further modulated by an oscillator 16, a multiplier 17, and a BPF 18 to generate a transmit signal to be sent to the center station 30. A terminal station 20 is also the same. At the center station 30, the arrival multiplexed transmit signal is demodulated by a BPF 31, an oscillator 32, and a multiplier 33, subchannel demodulation is performed by an AD converter 34, an SP converter 35, and an FFT 36, and the symbol sequences sent from the respective terminal stations are decoded by a DEMUX 37 and PS converters 38a and 38b.

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11/00			11/00	Z	

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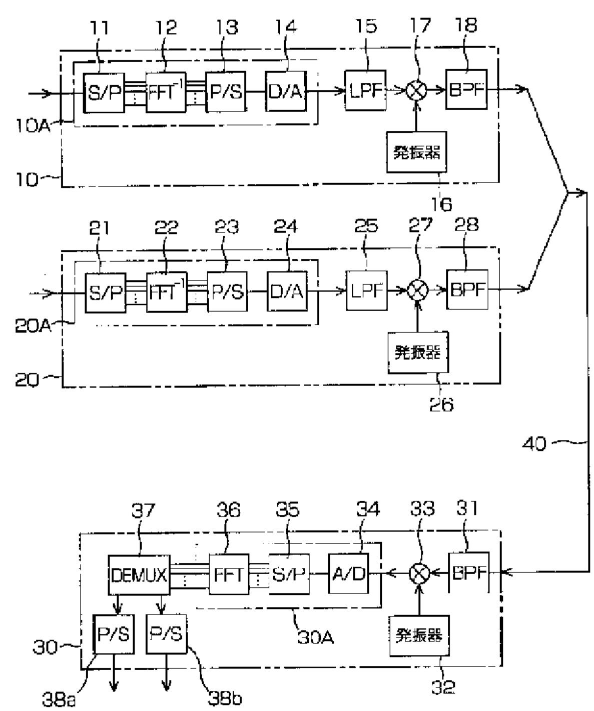
(54)【発明の名称】　デジタル伝送システム

(57)【要約】

【課題】　端末局の増加に容易に対処することが可能で、且つ、周波数帯域を有効に利用しつつモノトーン雑音に強いデジタル伝送システムを提供する。

【解決手段】　端末局１０において、ＳＰ変換器１１、逆ＦＦＴ１２、ＰＳ変換器１３およびＤＡ変換器１４により、端末局に固有の複数のサブチャンネルキャリア信号がセンタ局３０へ送信すべきシンボル列により変調され、さらに、発振器１６、乗算器１７およびＢＰＦ１８により、キャリア信号が変調されて、センタ局３０に送信されるべき伝送信号となる。端末局２０においても同様である。センタ局３０においては、ＢＰＦ３１、発振器３２および乗算器３３により、到達した合波された伝送信号が復調され、ＡＤ変換器３４、ＳＰ変換器３５およびＦＦＴ３６によりサブチャンネル復調され、ＤＥＭＵＸ３７およびＰＳ変換器３８ａ，３８ｂにより、各端末局から送信されたシンボル列が復元される。



【特許請求の範囲】

【請求項１】　センタ局と２以上の所定数の端末局それぞれとの間におけるマルチキャリア伝送方式によるデジタル伝送システムであって、
前記所定数の端末局それぞれに設けられ、各端末局に固有の複数のサブチャンネルキャリア信号を前記センタ局に送信すべきシンボル列で変調してベースバンド時系列信号を生成し出力するサブチャンネル変調手段と、
前記所定数の端末局それぞれに設けられ、前記所定数の端末局すべてに共通のキャリア信号を前記ベースバンド時系列信号で変調して伝送信号を出力する変調手段と、
前記所定数の端末局それぞれで生成された前記伝送信号を合波して合波信号とし、該合波信号をセンタ局に向けて送出する合波手段と、
前記センタ局に設けられ、到達した前記合波信号を前記キャリア信号について復調して、ベースバンド時系列混成信号を出力する復調手段と、
前記センタ局に設けられ、前記ベースバンド時系列混成信号を前記所定数の端末局すべての前記複数のサブチャンネルキャリア信号それぞれについて復調して、シンボル列混成信号を出力するサブチャンネル復調手段と、
前記センタ局に設けられ、前記シンボル列混成信号を前記所定数の端末局それぞれから送信されたシンボル列それぞれに分離して出力する分離手段と、
前記所定数の端末局それぞれから前記シンボル列が送信される送信速度を調整するシンボルレート調整手段と、
前記所定数の端末局それぞれから前記伝送信号それぞれを送出するタイミングを調整して、前記所定数の端末局それぞれから送出された前記伝送信号それぞれが前記センタ局に到達する時刻を一定にする送出タイミング調整手段と、
を備えることを特徴とするデジタル伝送システム。

【請求項２】　前記所定数の端末局それぞれに固有の前記複数のサブチャンネルキャリア信号それぞれの周波数は基準周波数の整数倍であり、
前記サブチャンネル変調手段は、
前記センタ局に送信すべきシンボル列をパラレル信号に変換する第１のシリアルーパラレル変換器と、
前記第１のシリアルーパラレル変換器からの出力信号を、前記複数のサブチャンネルキャリア信号それぞれの周波数について逆フーリエ変換する逆フーリエ変換器と、
前記逆フーリエ変換器からの出力信号をシリアル信号に変換する第１のパラレルーシリアル変換器と、
前記第１のパラレルーシリアル変換器からの出力信号をアナログ信号に変換して前記ベースバンド時系列信号を出力するデジタルーアナログ変換器と、
を備え、
前記サブチャンネル復調手段は、
前記ベースバンド時系列混成信号をデジタル信号に変換するアナログーデジタル変換器と、
前記アナログーデジタル変換器からの出力信号をパラレル信号に変換する第２のシリアルーパラレル変換器と、
前記第２のシリアルーパラレル変換器からの出力信号を、前記所定数の端末局それぞれの前記複数のサブチャンネルキャリア信号それぞれの周波数についてフーリエ変換して前記シンボル列混成信号を出力するフーリエ変換器と、
を備え、
前記分離手段は、
前記シンボル列混成信号を前記所定数の端末局それぞれから送信されたシンボル列それぞれに対応するパラレル信号それぞれに分離するデマルチプレクサと、
前記デマルチプレクサで分離されたパラレル信号それぞれをシリアル信号に変換して出力する第２のパラレルーシリアル変換器と、
を備える、
ことを特徴とする請求項１記載のデジタル伝送システム。

【請求項３】　前記所定数の端末局それぞれに固有の前記複数のサブチャンネルキャリア信号それぞれの周波数が一定の周波数帯域内で混在している、ことを特徴とする請求項１記載のデジタル伝送システム。

【請求項４】　前記所定数の端末局それぞれに設けられ、強度制御信号に基づいて前記伝送信号の強度を調整する伝送信号強度調整手段と、
前記センタ局に到達した前記合波信号の強度に基づいて、前記所定数の端末局それぞれから送出された前記伝送信号それぞれの強度レベルを求め、該強度レベルに基づいて前記強度制御信号それぞれを生成して対応する端末局それぞれに送出する伝送信号強度制御手段と、
を更に備えることを特徴とする請求項１記載のデジタル伝送システム。

【発明の詳細な説明】

【０００１】
【発明の属する技術分野】本発明は、例えばＣＡＴＶ伝送路を利用した樹枝状ネットワークにおいて好適に用いられるデジタル信号伝送技術に関するものである。

【０００２】
【従来の技術】従来よりＣＡＴＶ伝送路網を利用した大容量伝送技術の研究・開発が進められている。例えば、ＣＡＴＶ基盤技術研究所編集の「研究開発報告書」（平成５年３月３０日）には、時分割多重（ＴＤＭＡ：Time Divison Multiple Access）方式を採用して、遅延計測劣化分析、抑圧安定化方法、実伝送路変動と伝送品質との関係などに関する研究を行った結果が報告されている。

【０００３】それによれば、所定の前提条件下で伝送品質上十分なマージンを確保しつつ最も多くの通話チャンネルを実現するという観点から、誤り訂正符号も符号化

変調も施さない４相位相変調（ＱＰＳＫ： Quadrature Phase Shift Keying）信号を遅延検波し、高精度遅延時間計測制御を施す方式がＣＡＴＶ網上でのＴＤＭＡシステムとして最適であると結論付けられている。

【０００４】この報告で最適とされているＴＤＭＡ方式は、ＱＰＳＫ１波を時分割で利用するものであり、キャリア周波数は特定の１波に予め決められている。また、各端末局のシンボル送出タイミングを制御すべく、センタ局から基準クロックが送出されている。キャリア周波数が１つであることから、センタ局においては受信装置が１式で済み、データ送信速度は或程度フレキシブルであるという特徴を有する。

【０００５】

【発明が解決しようとする課題】しかしながら、上記従来例では、各端末局のシンボル送出タイミングの条件が厳しく、センタ局による制御が容易でないという問題点がある。また、キャリア周波数が１つであり、また、誤り訂正符号も施さないため、モノトーン雑音が存在する場合に、全データがその雑音の為に受信不能に陥るという問題点もある。

【０００６】ところで、多重化技術にはＴＤＭＡの他に周波数分割多重（ＦＤＭ： Frequency Division Multiplexing）方式がある。中でも特に、ＦＤＭ方式の１種である直交周波数分割多重（ＯＦＤＭ： Orthogonal Frequency Division Multiplexing ）方式は、送信すべき情報で複数のサブチャンネルキャリア信号を変調してベースバンド時系列信号とし、更に、このベースバンド時系列信号で１つのキャリア信号を変調して、その結果を相手局に送信するマルチキャリア伝送方式である。このマルチキャリア伝送方式は、多数のサブチャンネルキャリア信号を用いることからゴーストのある伝送路での周波数選択制フェーディングに強い、誤り訂正符号化の効果が大きい、周波数帯域の利用効率が高い、等の多くの利点を有し、前述のＴＤＭＡ方式の問題を解決するものである（例えば、テレビジョン学会誌 Vol.50, No.1, pp.24-41 (1996)）。

【０００７】しかし、ＣＡＴＶ伝送路網のように１つのセンタ局と複数の端末局それぞれとの間でデータ伝送を行う樹枝状ネットワークに、このマルチキャリア伝送方式をそのまま適用する場合には、以下のような問題点がある。すなわち、端末局毎に異なるキャリア信号を用いる必要があることから、センタ局においては、端末局の個数と同数またはそれ以上の復調器等が必要となり、システムが大規模になる。それだけでなく、センタ局に設けられた復調器の個数を越えて端末局が増加した場合に、センタ局では直ちには対処できないという問題点もある。

【０００８】また、端末局からセンタ局に送信する場合に誤り訂正符号を施せばランダム誤りに対しては訂正可能であるが、モノトーン雑音により生じたバースト誤りが１つの伝送チャンネル（１つの端末局からのキャリア信号）に集中すると、その端末局からの伝送信号についてはセンタ局で誤り訂正復号化することができないという問題点もある。

【０００９】本発明は、上記問題点を解消する為になされたものであり、複数の端末局と１つのセンタ局との間の樹枝状ネットワークのＣＡＴＶ伝送路網を利用する場合であっても、端末局の増加に容易に対処することが可能で、且つ、周波数帯域を有効に利用しつつモノトーン雑音に強いデジタル伝送システムを提供することを目的とする。

【００１０】

【課題を解決するための手段】本発明に係るデジタル伝送システムは、センタ局と２以上の所定数の端末局それぞれとの間におけるマルチキャリア伝送方式によるデジタル伝送システムであって、(1) 所定数の端末局それぞれに設けられ、各端末局に固有の複数のサブチャンネルキャリア信号をセンタ局に送信すべきシンボル列で変調してベースバンド時系列信号を生成し出力するサブチャンネル変調手段と、(2) 所定数の端末局それぞれに設けられ、所定数の端末局すべてに共通のキャリア信号をベースバンド時系列信号で変調して伝送信号を出力する変調手段と、(3) 所定数の端末局それぞれで生成された伝送信号を合波して合波信号とし、この合波信号をセンタ局に向けて送出する合波手段と、(4) センタ局に設けられ、到達した合波信号をキャリア信号について復調して、ベースバンド時系列混成信号を出力する復調手段と、(5) センタ局に設けられ、ベースバンド時系列混成信号を所定数の端末局すべての複数のサブチャンネルキャリア信号それぞれについて復調して、シンボル列混成信号を出力するサブチャンネル復調手段と、(6) センタ局に設けられ、シンボル列混成信号を所定数の端末局それぞれから送信されたシンボル列それぞれに分離して出力する分離手段と、(7) 所定数の端末局それぞれからシンボル列が送信される送信速度を調整するシンボルレート調整手段と、(8) 所定数の端末局それぞれから伝送信号それぞれを送出するタイミングを調整して、所定数の端末局それぞれから送出された伝送信号それぞれがセンタ局に到達する時刻を一定にする送出タイミング調整手段と、を備えることを特徴とする。

【００１１】このデジタル伝送システムにおいては、所定数の端末局それぞれにおいて、サブチャンネル変調手段により、各端末局に固有の複数のサブチャンネルキャリア信号はセンタ局に送信すべきシンボル列で変調されてベースバンド時系列信号とされ、変調手段により、所定数の端末局すべてに共通のキャリア信号はベースバンド時系列信号で変調されて伝送信号が出力される。所定数の端末局それぞれで生成された伝送信号は、合波手段により合波されて合波信号となり、この合波信号はセンタ局に向けて送出される。

【００１２】センタ局においては、到達した合波信号は、復調手段により一括してキャリア信号について復調されてベースバンド時系列混成信号となり、そのベースバンド時系列混成信号は、サブチャンネル復調手段により一括して所定数の端末局すべての複数のサブチャンネルキャリア信号それぞれについて復調されてシンボル列混成信号となり、そのシンボル列混成信号は、分離手段により所定数の端末局それぞれから送信されたシンボル列それぞれに分離される。

【００１３】これに際して、所定数の端末局それぞれからシンボル列が送信される送信速度は、シンボルレート調整手段により調整され、センタ局に到達するデータ量がセンタ局の受信能力を越えることはない。また、所定数の端末局それぞれから伝送信号それぞれを送出するタイミングは、送出タイミング調整手段により調整されて、所定数の端末局それぞれから送出された伝送信号それぞれはセンタ局に同時に到達する。

【００１４】所定数の端末局それぞれに固有の複数のサブチャンネルキャリア信号それぞれの周波数は基準周波数の整数倍であり、サブチャンネル変調手段は、(1-a) センタ局に送信すべきシンボル列をパラレル信号に変換する第１のシリアル－パラレル変換器と、(1-b) 第１のシリアル－パラレル変換器からの出力信号を、複数のサブチャンネルキャリア信号それぞれの周波数について逆フーリエ変換する逆フーリエ変換器と、(1-c) 逆フーリエ変換器からの出力信号をシリアル信号に変換する第１のパラレル－シリアル変換器と、(1-d) 第１のパラレル－シリアル変換器からの出力信号をアナログ信号に変換してベースバンド時系列信号を出力するデジタル－アナログ変換器と、を備え、サブチャンネル復調手段は、(2-a) ベースバンド時系列混成信号をデジタル信号に変換するアナログ－デジタル変換器と、(2-b) アナログ－デジタル変換器からの出力信号をパラレル信号に変換する第２のシリアル－パラレル変換器と、(2-c) 第２のシリアル－パラレル変換器からの出力信号を、所定数の端末局それぞれの複数のサブチャンネルキャリア信号それぞれの周波数についてフーリエ変換してシンボル列混成信号を出力するフーリエ変換器と、を備え、分離手段は、(3-a) シンボル列混成信号を所定数の端末局それぞれから送信されたシンボル列それぞれに対応するパラレル信号それぞれに分離するデマルチプレクサと、(3-b) デマルチプレクサで分離されたパラレル信号それぞれをシリアル信号に変換して出力する第２のパラレル－シリアル変換器と、を備えるものでもよい。この場合には、ＯＦＤＭ方式に準じた方式で、シンボル列は各端末局からセンタ局に伝送される。

【００１５】所定数の端末局それぞれに固有の複数のサブチャンネルキャリア信号それぞれの周波数が一定の周波数帯域内で混在している場合には、誤り訂正符号化技術を併用することにより、ランダム誤りだけでなく、モノトーン雑音などにより生じるバースト誤りにも強い伝送が実現できる。

【００１６】本発明に係るデジタル伝送システムは、更に、(1) 所定数の端末局それぞれに設けられ、強度制御信号に基づいて伝送信号の強度を調整する伝送信号強度調整手段と、(2) センタ局に到達した合波信号の強度に基づいて、所定数の端末局それぞれから送出された伝送信号それぞれの強度レベルを求め、その強度レベルに基づいて強度制御信号それぞれを生成して対応する端末局それぞれに送出する伝送信号強度制御手段と、を備えてもよい。この場合、伝送信号強度制御手段により出力された強度制御信号に基づいて、伝送信号強度調整手段により各端末局それぞれからセンタ局に伝送される伝送信号それぞれが互いに略等しい強度になるので、雑音に強い伝送が可能となる。

【００１７】

【発明の実施の形態】以下、添付図面を参照して本発明の実施の形態を詳細に説明する。尚、図面の説明において同一の要素には同一の符号を付し、重複する説明を省略する。なお、以下では、簡便のため端末局が２つである場合を想定して説明する。図１は、本実施形態に係るデジタル伝送システムの構成図である。

【００１８】本実施形態に係るデジタル伝送システムは、端末局１０，２０それぞれとセンタ局３０との間でデジタル伝送を行うものである。また、端末局１０および端末局２０それぞれには同様の装置が備えられている。したがって、一方の端末局１０について主に説明する。また、マルチキャリア伝送方式としてＯＦＤＭ方式に準じた伝送方式を採用した場合について説明する。

【００１９】端末局１０には、その端末局１０に固有のサブチャンネルキャリアをセンタ局３０に送信すべきシンボル列で変調してベースバンド時系列信号を生成するサブチャンネル変調手段１０Ａとして、シリアル－パラレル変換器（以下、ＳＰ変換器）１１、逆フーリエ変換器（以下、逆ＦＦＴ）１２、パラレル－シリアル変換器（以下、ＰＳ変換器）１３、および、デジタル－アナログ変換器（以下、ＤＡ変換器）１４が備えられている。また、端末局１０には、サブチャンネル変調手段１０Ａから出力されたベースバンド時系列信号でキャリア信号を変調して伝送信号を出力する変調手段として、発振器１６、乗算器１７およびバンドパスフィルタ（以下、ＢＰＦ）１８が備えられている。

【００２０】端末局１０からセンタ局３０へ送信すべきシンボル列（シリアルデータ）は、先ず、ＳＰ変換器１１により所定データ長のパラレルデータに変換される。そして、パラレルデータとされたシンボル列は、逆ＦＦＴ１２に入力され逆フーリエ変換される。すなわち、パラレルデータとされたシンボル列d_k(k=1, 2, 3, …) は、

【数１】

$$x(n\cdot\Delta T)=\sum_{k=0}^{N-1}\{d_k\cdot\exp(j2\pi nk/N)\} \quad ---- \quad (1)$$

なる変換式に従って、サブチャンネルキャリア変調信号x（n・ΔT）に変換される。ここで、ΔTはサンプリング間隔、n・ΔTはサンプリング点、ｊは虚数単位、πは円周率、Nはサブチャンネルキャリア信号の個数である。また、サブチャンネルキャリア信号は、

【数２】

$$\exp(j2\pi nk/N) \quad ---- \quad (2)$$

である。（1）式および（2）式から解るように、サブチャンネルキャリア信号がシンボル列で変調された形となっている。このような機能は、ＤＳＰ（Digital Signal Processor）を用いて容易に実現することができる。

【００２１】この（２）式で表されるサブチャンネルキャリア信号の波形は、直交性を有している。すなわち、互いに等しいｎ値を有する２つのサブチャンネルキャリア信号の積を１周期に亘って時間積分すると０でない有限値となるが、互いに異なるｎ値を有する２つのサブチャンネルキャリア信号の積を１周期に亘って時間積分すると０になる。

【００２２】このようにして逆ＦＦＴ１２により逆フーリエ変換されたシンボル列は、ＰＳ変換器１３により再びシリアルデータに変換され、ＤＡ変換器１４によりアナログデータに変換されて、ベースバンド時系列信号となる。このベースバンド時系列信号は、ローパスフィルタ（以下、ＬＰＦ）１５により高周波数成分がカットされ、発振器１６から出力されたキャリア信号と乗算器１７により乗算され、そして、ＢＰＦ１８で所定の帯域の周波数成分の信号のみが通過する。このようにベースバンド時系列信号でキャリア信号が変調されて、センタ局３０に送信されるべき伝送信号となる。

【００２３】同様に、端末局２０では、センタ局３０へ送信すべきシンボル列（シリアルデータ）は、ＳＰ変換器２１によりパラレルデータに変換され、逆ＦＦＴ２２により逆フーリエ変換され、ＰＳ変換器２３により再びシリアルデータに変換され、ＤＡ変換器２４によりアナログデータに変換されて、ベースバンド時系列信号となる。このベースバンド時系列信号は、ＬＰＦ２５により高周波数成分がカットされ、発振器２６から出力されたキャリア信号と乗算器２７により乗算され、そして、ＢＰＦ２８で所定の帯域の周波数成分の信号のみが通過して、センタ局３０に送信されるべき伝送信号となる。ここで、キャリア信号は、端末局２０において発振器２６から出力されるキャリア信号と同一周波数である。

【００２４】このようにして端末局１０および２０それぞれで生成された伝送信号は、合波されてセンタ局３０に送信される。これに際して、それぞれの伝送信号は以下の条件を満足する必要がある。

【００２５】第１に、端末局１０において用いられるサブチャンネルキャリア信号の周波数 f_k(k=1, 2, 3, …) と、端末局２０において用いられるサブチャンネルキャリア信号の周波数 g_k(k=1, 2, 3, …) とは、同一のものが存在しないことが必要である。これは、端末局１０および２０それぞれからの伝送信号が合波されてセンタ局３０に到達したときに、これらがセンタ局３０で分離可能でなければならないからである。したがって、例えば、図２に示すように、端末局１０において用いられるサブチャンネルキャリア信号の周波数 f_k(k=1, 2, 3, …) と端末局２０において用いられるサブチャンネルキャリア信号の周波数 g_k(k=1, 2, 3, …) とを交互に且つ等間隔に並ぶ値とする。すなわち、周波数 f_k(k=1, 2, 3, …) を、

【数３】

$$f_k=(2k-1)/(N\cdot\Delta T) \quad ---- \quad (3)$$

とし（図２（ａ））、周波数 g_k(k=1, 2, 3, …) を、

【数４】

$$g_k=2k/(N\cdot\Delta T) \quad ---- \quad (4)$$

とする（図２（ｂ））。このようにすれば、これらが合波された伝送信号は、後に説明するようにセンタ局３０において一括して復調した後に分離することができる。

【００２６】第２に、端末局１０および２０それぞれから単位時間あたりに送出されるシンボル列それぞれの長さ（シンボルレート）が互いに等しいことが必要である。このために、例えば、各端末局におけるシンボルレートを予め固定的に設定しておいてもよい。また、センタ局３０より端末局１０および２０それぞれに対して、シンボルレート情報を送信し、端末局１０および２０それぞれは、そのシンボルレート情報に従った単位時間当たりのデータ長のシンボル列をセンタ局３０に送出するようにしてもよい。このようにすれば、ネットワークにつながる端末局の個数が増えたとき、或いは、センタ局３０に送信している端末局の個数が増えたときには、センタ局３０は、各端末局に対してシンボルレートを小さくするようシンボルレート情報により指示することにより、センタ局３０に到達するデータ量がセンタ局３０の処理能力を越えないようにすることができる。

【００２７】第３に、端末局１０および２０それぞれから送出された伝送信号がセンタ局３０に同時刻に到着することが必要である。これは、後に説明するように、端末局１０および２０それぞれから送出され合波された伝送信号をセンタ局３０で一括して復調するためである。したがって、端末局１０および２０それぞれとセンタ局３０との間の遅延時間を計測し、これに基づいて、端末局１０および２０それぞれからの伝送信号の送出タイミングを調整する。

【００２８】具体的には、例えば、センタ局３０から端末局１０に所定の信号を送信し、端末局１０はその所定の信号を受信して今度はセンタ局３０に向けて所定の信号を送信し、そして、センタ局３０は端末局１０から到達した所定の信号を受信して、センタ局３０はこの間の時間を計測し、この時間に対応する遅延時間を端末局１

０に指示する。あるいは、ＣＡＴＶ網の場合には、端末局１０とセンタ局３０との間の線路の長さが判れば、この間の遅延時間が判るので、この線路の長さに対応する遅延時間を端末局１０に予め設定しておいてもよい。端末局１０における遅延時間の設定は、例えば、逆ＦＦＴ１２の後段にレジスタ（図示せず）を設けて、逆ＦＦＴ１２の出力データを所定時間の間ホールドする。あるいは、ＰＳ変換器１３の後段にＦＩＦＯメモリ（図示せず）を設けて、ＰＳ変換器１３の出力データを所定時間の間だけ遅延させて出力するようにしてもよい。端末局２０についても同様である。

【００２９】以上のようにして端末局１０および２０それぞれから出力された伝送信号それぞれは樹枝状ネットワークの伝送路４０に送出されて合波され、その合波された伝送信号はセンタ局３０に到達する。このセンタ局３０には、合波されて到達した伝送信号を一括して復調してベースバンド時系列混成信号を出力する復調手段として、ＢＰＦ３１、発振器３２および乗算器３３が備えられている。また、センタ局３０には、ベースバンド時系列混成信号を復調して各端末局から送信されてきたシンボル列混成信号を出力するサブチャンネル復調手段３０Ａとして、アナログーデジタル変換器（以下、ＡＤ変換器）３４、ＳＰ変換器３５およびフーリエ変換器（以下、ＦＦＴ）３６が備えられている。また、センタ局３０には、シンボル列混成信号を各端末局それぞれから送信されてきたシンボル列それぞれに分離する分離手段として、デマルチプレクサ（以下、ＤＥＭＵＸ）３７、ＰＳ変換器３８ａ，３８ｂが備えられている。

【００３０】センタ局３０に到達した合波された伝送信号は、まず、ＢＰＦ３１で所定の帯域の周波数成分の信号のみが通過して、発振器３２から出力されたキャリア信号と乗算器３３により乗算されて復調され、ベースバンド時系列混成信号が出力される。このキャリア信号は、端末局１０および２０それぞれの発振器１６およに２６それぞれから出力されるキャリア信号と同じ周波数のものである。ベースバンド時系列混成信号は、端末局１０のサブチャンネル変調手段１０Ａで生成されたベースバンド時系列信号と、端末局２０のサブチャンネル変調手段２０Ａで生成されたベースバンド時系列信号とが混成された信号（図２（ｃ））である。

【００３１】このベースバンド時系列混成信号は、サブチャンネル復調手段３０Ａによりサブチャンネルキャリア毎に復調される。すなわち、ベースバンド時系列混成信号は、ＡＤ変換器３４によりアナログデータに変換され、ＳＰ変換器３５によりパラレルデータに変換され、ＦＦＴ３６によりフーリエ変換され、シンボル列混成信号が出力される。ここで、ＦＦＴ３６においてなされる演算は、（１）式に対応するフーリエ変換であって、

【数５】

$$d_k=\sum_{n=0}^{N-1}\{x(n\cdot\Delta T)\cdot\exp(-j2\pi nk/N)\}\quad ----\quad (5)$$

で表される。このＦＦＴ３６も、ＤＳＰにより容易に実現することができる。このシンボル列混成信号は、端末局１０から送信されたシンボル列と端末局２０から送信されたシンボル列とが混成された信号である。

【００３２】このシンボル列混成信号は、ＤＥＭＵＸ３７により、端末局１０および２０それぞれから送信されたシンボル列それぞれに分離される。このＤＥＭＵＸ３７は、シンボル列混成信号中の各シンボルが、どのサブチャンネルキャリア信号により送信されてきたかを判断し、これに基づいて端末局１０から送信されたシンボル列と端末局２０から送信されたシンボル列とに分離する。そして、端末局１０から送信されたシンボル列はＰＳ変換器３８ａによりシリアルデータに変換されて出力され、端末局２０から送信されたシンボル列はＰＳ変換器３８ｂによりシリアルデータに変換されて出力される。以上のようにして、端末局１０および２０それぞれから送信されたシンボル列それぞれは、センタ局３０で受信される。

【００３３】なお、センタ局３０に到達する伝送信号の強度が、その伝送信号が送出された端末局に依って異なる場合には、これを略一定レベルとすべく、伝送信号の強度を制御・調整する手段を備えていてもよい。例えば、センタ局３０において、ベースバンド時系列混成信号をサブチャンネルキャリア信号毎にモニタして、端末局１０および２０それぞれから送出された伝送信号の強度を求め、これが略一定となるように端末局１０および２０それぞれにフィードバックして制御する。端末局１０および２０それぞれにおいては、ＤＡ変換器１４および２４それぞれの後段に増幅器あるいは減衰器を設けて、センタ局３０からの指示に従い伝送路４０に送出する伝送信号の強度レベルを調整する。このようにすることにより、更に雑音に強いデジタル伝送システムを実現することができる。

【００３４】以上のように本システムは、マルチキャリア方式を採用するとともに、サブチャンネルキャリア信号の周波数を各端末局に固有のものとし、且つ、キャリア信号の周波数を全ての端末局に共通のものとしたので、１つのキャリア信号で全ての端末局からのシンボル列をセンタ局に送信することができ、しかも、センタ局では復調手段およびサブチャンネル復調手段が一式で済むため、システム全体の構成が簡単で安価となる。また、ネットワークにつながる端末局の個数が増加したとき、あるいは、センタ局へ同時に送信している端末局が増加したときにも、センタ局からの指示により各端末局のシンボルレートを小さくし、各端末局の逆ＦＦＴが扱うデータサイズを小さくすることにより、容易に対処することができる。

【００３５】また、一方の端末局のサブチャンネルキャ

リア信号の周波数と他方の端末局のサブチャンネルキャリア信号の周波数とを交互に設定することとしたので、特にＣＡＴＶ伝送網で問題となる流合雑音や無線による妨害波が伝送路に混入した場合でも、それらの雑音の影響は、或る端末局から送出された伝送信号に集中することなく、多数の端末局から送出された伝送信号それぞれに分散する。したがって、個々の端末局から送出された伝送信号にとってはランダム誤りとなるので、誤り訂正符号化を施しておけば誤り訂正を行うことが可能である。さらに、時間インターリーブあるいは周波数インターリーブと誤り訂正符号化とを併用すれば、更に、バースト誤りに強い伝送を実現することができる。なお、雑音が問題とならないような伝送路においては、各端末局に固有のサブチャンネルキャリア信号の周波数は、図２（ｃ）に示すように端末局それぞれに対応するものが交互に並んでいる必要はなく、端末局毎にまとまっていても構わない。

【００３６】本発明は複数点映像収集システムに応用することができる。例えば、１つの伝送チャンネルが２０Ｍｂｐｓである場合に、３つの端末局それぞれからの映像に6Ｍｂｐｓを割り当てて精細な映像を伝送することとし、一方、他の１つの端末局からの映像については粗い映像で構わない場合には2Ｍｂｐｓを割り当ててデータ量の少ない映像を伝送することができる。このように、目的に応じてデータ量の異なる映像をセンタ局に送信することができる。

【００３７】本発明は、上記実施形態に限定されるものではなく種々の変形が可能である。例えば、マルチキャリア伝送方式としては、ＯＦＤＭ方式に限られるものではなく、他のマルチキャリア伝送方式、例えばＤＭＴ（Digital Multi-tone）方式に準じた伝送方式であっても構わない。端末局の数は２に限られるものではなく、３以上であっても構わない。また、伝送路網はＣＡＴＶに限るものではなく、他の樹枝状ネットワークの伝送路網においても適用可能である。

【００３８】また、各端末局のサブチャンネルキャリア信号の周波数の設定は等間隔に限られるものではない。１の端末局のサブチャンネルキャリア信号の周波数と他の端末局のサブチャンネルキャリア信号の周波数とは交互に等間隔に設定されていなくても構わない。相互に直交性を有し、且つ、異なる端末局の間で同一の周波数を設定することのない限りにおいて、任意のサブチャンネルキャリア信号の周波数を用いることができる。

【００３９】また、各端末局およびセンタ局それぞれのキャリア信号の周波数は、予め設定しておいて固定してもよいし、センタ局からの指示により各端末局の発振器を制御してキャリア信号の周波数を変更するようにしてもよい。この場合、発振器として電圧制御可能な水晶発振器（ＶＣＸＯ）が好適に用いられる。また、各端末局のサブチャンネル変調手段におけるサブチャンネルキャリア信号の周波数も、予め設定して固定してもよいし、センタ局からの指示により各端末局の逆ＦＦＴにおける演算パラメタを変更してサブチャンネルキャリア信号の周波数を変更するようにしてもよい。

【００４０】

【発明の効果】以上、詳細に説明したとおり本発明によれば、各端末局それぞれにおいて、各端末局に固有の複数のサブチャンネルキャリア信号がセンタ局に送信すべきシンボル列で変調されてベースバンド時系列信号とされ、更に、端末局すべてに共通のキャリア信号がベースバンド時系列信号で変調されて伝送信号とされ、各端末局それぞれで生成された伝送信号が合波されて合波信号となり、この合波信号がセンタ局に向けて送出される。

【００４１】センタ局においては、到達した合波信号は、復調手段により、一括してキャリア信号について復調されてベースバンド時系列混成信号となり、更に、そのベースバンド時系列混成信号は、サブチャンネル復調手段により、一括して所定数の端末局すべての複数のサブチャンネルキャリア信号それぞれについて復調されてシンボル列混成信号となり、そのシンボル列混成信号は、分離手段により所定数の端末局それぞれから送信されたシンボル列それぞれに分離される。

【００４２】これに際して、所定数の端末局それぞれから送信されるシンボル列の送信速度は調整され、センタ局に到達するデータ量がセンタ局の受信能力を越えることはない。また、所定数の端末局それぞれから伝送信号を送出するタイミングは調整されて、所定数の端末局それぞれから送出された伝送信号それぞれはセンタ局に同時に到達する。

【００４３】このような構成としたことにより、キャリア信号の周波数は１つだけであるので、センタ局においては、復調手段およびサブチャンネル復調手段は１式のみで済み、構成が簡単となり安価となる。センタ局につながる端末局が増えた場合、或いは、センタ局へ同時に伝送する端末局が増えた場合であっても、新たな装置を付加することなく、これらの事態に容易に対処可能である。

【００４４】また、各端末局それぞれに固有の複数のサブチャンネルキャリア信号それぞれの周波数が一定の周波数帯域内で混在している場合には、誤り訂正符号化技術を併用することにより、ランダム誤りだけでなく、モノトーン雑音などにより生じるバースト誤りにも強い伝送が実現できる。

【００４５】また、各端末局それぞれから送出される伝送信号の強度が互いに略等しい強度になるよう調整すれば、更に雑音に強い伝送が可能となる。

【図面の簡単な説明】

【図１】本実施形態に係るデジタル伝送システムの構成図である。

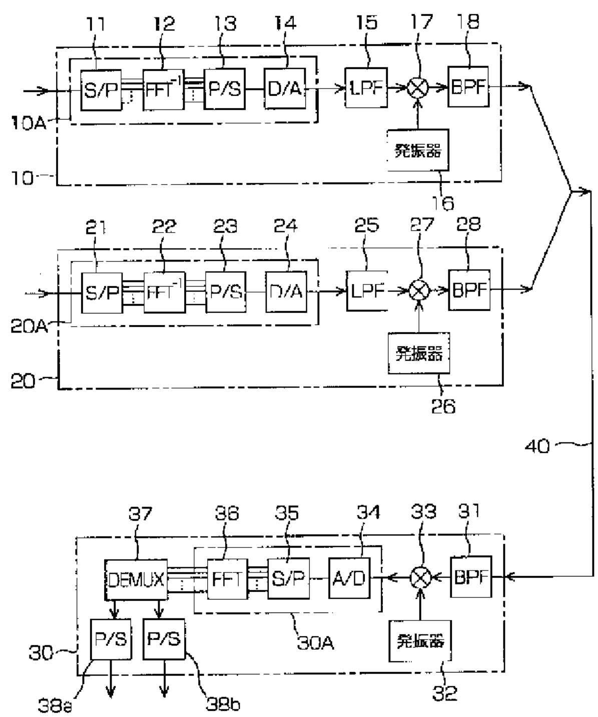
【図２】サブチャンネルキャリア信号の周波数の説明図

である。

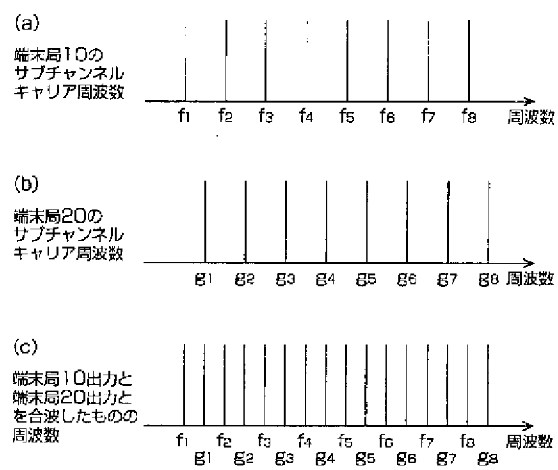
【符号の説明】

１０…端末局、１０Ａ…サブチャンネル変調手段、１１…シリアル－パラレル変換器（ＳＰ変換器）、１２…逆フーリエ変換器（逆ＦＦＴ）、１３…パラレル－シリアル変換器（ＰＳ変換器）、１４…デジタル－アナログ変換器（ＤＡ変換器）、１５…ローパスフィルタ（ＬＰＦ）、１６…発振器、１７…乗算器、１８…バンドパスフィルタ（ＢＰＦ）、２０…端末局、２０Ａ…サブチャンネル変調手段、２１…シリアル－パラレル変換器（ＳＰ変換器）、２２…逆フーリエ変換器（逆ＦＦＴ）、２３…パラレル－シリアル変換器（ＰＳ変換器）、２４…デジタル－アナログ変換器（ＤＡ変換器）、２５…ローパスフィルタ（ＬＰＦ）、２６…発振器、２７…乗算器、２８…バンドパスフィルタ（ＢＰＦ）、３０…センタ局、３０Ａ…サブチャンネル復調手段、３１…バンドパスフィルタ（ＢＰＦ）、３２…発振器、３３…乗算器、３４…アナログ－デジタル変換器（ＡＤ変換器）、３５…シリアル－パラレル変換器（ＳＰ変換器）、３６…フーリエ変換器（ＦＦＴ）、３７…デマルチプレクサ（ＤＥＭＵＸ）、３８ａ，３８ｂ…パラレル－シリアル変換器（ＰＳ変換器）、４０…伝送路。

【図１】



【図２】



JP9327073A METHOD FOR ARRANGING AND TRANSMITTING PILOT CHANNEL IN CDMA MOBILE COMMUNICATION SYSTEM

Bibliography

DWPI Title

Pilot channel assignment method in CDMA mobile communication system involves performing time sharing of radio channel to number of time slots in direction of mobile station

Original Title

METHOD FOR ARRANGING AND TRANSMITTING PILOT CHANNEL IN CDMA MOBILE COMMUNICATION SYSTEM

Assignee/Applicant

Standardized: **NIPPON TELEGRAPH & TELEPHONE**
Original: N T T IDO TSUSHINMO KK

Inventor

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Publication Date (Kind Code)

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Application Number / Date

JP1996145910A / 1996-06-07

Priority Number / Date / Country

JP1996145910A / 1996-06-07 / JP

Abstract

PROBLEM TO BE SOLVED: To reduce the influence of a pilot channel on the diffusion code shortage by time-dividing and multiplexing an outgoing radio channel into plural time slots and assigning one of them as a pilot channel.

SOLUTION: The whole outgoing radio channels consisting of X-number diffusion codes have frame configuration and one frame is divided into the plural time slots so as to be multiplexed by time division. Then, the specified time slot of the radio channel diffused by the specified diffusion code is assigned as the pilot channel. The other time slots and the other diffusion codes are used as the communication channel for communication with the mobile station. For example, time slot numbers 1-4 are given to the four time slots in the frame in order from an early one in terms of time, the time slot #1 of the radio channel where the diffusion code is diffused by one is assigned as the pilot channel and the other time slots and the diffusion codes are assigned as the communication channel.

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Ｈ０４Ｂ　7/26				Ｐ	

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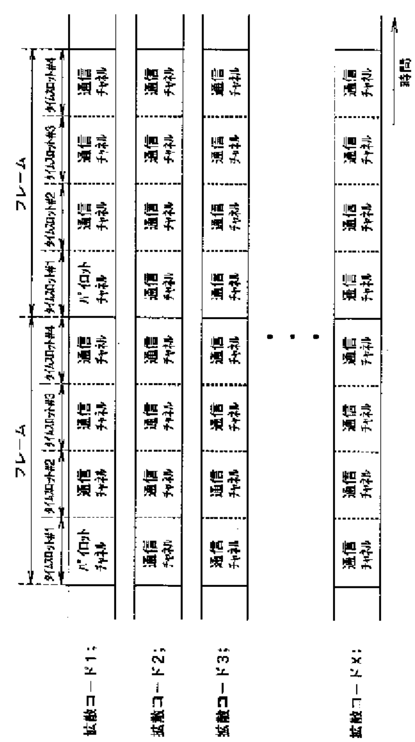
(54)【発明の名称】　ＣＤＭＡ移動通信システムにおけるパイロットチャネル配置および送信方法

(57)【要約】

【課題】　下り無線チャネルを時間分割多重して、パイロットチャネルを効率的に割り当てるとともに、他セルに対するパイロットチャネルの干渉電力を低減し得るＣＤＭＡ移動通信システムにおけるパイロットチャネル配置および送信方法を提供する。

【解決手段】　下り無線チャネルを複数のタイムスロットに時間分割して時間分割多重し、複数のタイムスロットのうちの１つをパイロットチャネルとして割り当てている。



【特許請求の範囲】

【請求項１】　複数のセルの各々に基地局が設けられ、各基地局は同一周波数で変調され、それぞれ異なって割り当てられた拡散コードで拡散されたパイロットチャネルを送信し、移動局は前記パイロットチャネルを受信することにより在圏セルを判定するＣＤＭＡ移動通信システムにおけるパイロットチャネル配置および送信方法であって、

基地局から移動局方向への下り無線チャネルを複数のタイムスロットに時間分割して時間分割多重し、前記複数のタイムスロットのうちの１つをパイロットチャネルとして割り当てることを特徴とするＣＤＭＡ移動通信システムにおけるパイロットチャネル配置および送信方法。

【請求項２】　パイロットチャネルとして割り当てる前記タイムスロットの時間的な位置を全セルで共通とすることを特徴とする請求項１記載のＣＤＭＡ移動通信システムにおけるパイロットチャネル配置および送信方法。

【請求項３】　パイロットチャネルとして割り当てる前記タイムスロットのみ常時一定の送信電力で送信し、移動局との通信用の他のタイムスロットは送信電力制御を行うことを特徴とする請求項１または２記載のＣＤＭＡ移動通信システムにおけるパイロットチャネル配置および送信方法。

【発明の詳細な説明】

【０００１】

【発明の属する技術分野】本発明は、符号分割多元接続方式（以下、ＣＤＭＡと略称する）の移動通信システムにおいて複数のセルの各々に基地局が設けられ、各基地局は同一周波数で変調され、それぞれ異なって割り当てられた拡散コードで拡散されたパイロットチャネルを送信し、移動局は前記パイロットチャネルを受信することにより在圏セルを判定するＣＤＭＡ移動通信システムにおけるパイロットチャネル配置および送信方法に関する。

【０００２】

【従来の技術】ＣＤＭＡ移動通信システムにおいて、下り無線チャネルでは、全ての移動機において同一セル内の他の複数移動局に対して送信された下り無線チャネルは全て干渉電力となり、受信品質を劣化させるかもしくは無線チャネル容量を劣化させる要因となる。しかし各下り無線チャネルに、お互いに直交化した複数の拡散コードを用い、基地局から同一拡散コード位相で送信することにより、同一セルで送信される他の下り無線チャネルは直交化されて干渉電力量を０にすることができる。ただしマルチパス環境では、拡散コード位相の異なる無線チャネルが発生するため、干渉電力量は０にはならないが、干渉電力を大きく低減することが可能である。これにより下り無線チャネルの受信品質の向上もしくは無線チャネル容量の向上を見込むことができる。

【０００３】図３に従来のパイロットチャネル配置方法を示す。従来は直交化した拡散コードの内の１つをパイロットチャネル用の拡散コードとして割り当て、この拡散コードで常時パイロットチャネルを送信し、他の拡散コードは移動局との通信に用いる通信チャネル用としていた。図３はＸ個ある直交化した拡散コードの内、拡散コード１をパイロットチャネル用に割り当てている場合を示している。

【０００４】更に移動局の在圏セル判定のために、隣接セルに在圏する移動局においても自局のパイロットチャネルを受信できるように、パイロットチャネルは移動局との通信に用いる無線チャネルより、比較的大きな送信電力で送信される必要がある。従来は前述の特定の拡散コードで、常時比較的大きな送信電力でパイロットチャネルは送信されていた。

【０００５】

【発明が解決しようとする課題】直交化した拡散コードの数はそれほど多くはない。本来ＣＤＭＡ移動通信システムでは無線チャネル容量は干渉電力量で決まるが、直交化した拡散コードを用いる場合、拡散コード数の不足により無線チャネル容量を十分に使用することができないという問題が起こりうる。このような状況において、前述したように１つの拡散コードをパイロットチャネルとして専有することにより、移動局との通信に用いる拡散コードがより不足するという問題があった。

【０００６】更に前述したように、パイロットチャネルは比較的大きな送信電力で常時送信されていたため、隣接セルに対して大きな干渉を与え、隣接セルの容量を減少させるという問題があった。

【０００７】本発明は、上記に鑑みてなされたもので、その目的とするところは、下り無線チャネルを時間分割多重して、パイロットチャネルを効率的に割り当てるとともに、他セルに対するパイロットチャネルの干渉電力を低減し得るＣＤＭＡ移動通信システムにおけるパイロットチャネル配置および送信方法を提供することにある。

【０００８】

【課題を解決するための手段】上記目的を達成するため、請求項１記載の本発明は、複数のセルの各々に基地局が設けられ、各基地局は同一周波数で変調され、それぞれ異なって割り当てられた拡散コードで拡散されたパイロットチャネルを送信し、移動局は前記パイロットチャネルを受信することにより在圏セルを判定するＣＤＭＡ移動通信システムにおけるパイロットチャネル配置および送信方法であって、基地局から移動局方向への下り無線チャネルを複数のタイムスロットに時間分割して時間分割多重し、前記複数のタイムスロットのうちの１つをパイロットチャネルとして割り当てることを要旨とする。

【０００９】また、請求項２記載の本発明は、請求項１記載の発明において、パイロットチャネルとして割り当

てる前記タイムスロットの時間的な位置を全セルで共通とすることを要旨とする。

【００１０】更に、請求項３記載の本発明は、請求項１または２記載の発明において、パイロットチャネルとして割り当てる前記タイムスロットのみ常時一定の送信電力で送信し、移動局との通信用の他のタイムスロットは送信電力制御を行うことを要旨とする。

【００１１】

【発明の実施の形態】以下、図面を用いて本発明の実施の形態について説明する。

【００１２】図１は、本発明の一実施形態に係るＣＤＭＡ移動通信システムにおけるパイロットチャネル配置および送信方法を実施するパイロットチャネルの配置方法の一例を示す図である。同図に示すように、Ｘ個の拡散コードからなる下り無線チャネルの全てはフレーム構成をとり、１フレームは複数のタイムスロットに分割され、時分割多重されている。図１では４個のタイムスロットに分割し、４多重した場合を示している。

【００１３】特定の拡散コードで拡散される無線チャネルの、特定のタイムスロットをパイロットチャネルとして割り当てる。他のタイムスロットおよび他の拡散コードは移動局との通信用の通信チャネルとして用いる。図１では、フレーム内の４個のタイムスロットについて時間的に早い順に１～４のタイムスロット番号をつけ、拡散コード１で拡散される無線チャネルのタイムスロット＃１をパイロットチャネルとして割り当て、他のタイムスロットおよび拡散コードを通信チャネルとして割り当てた場合を示している。これにより従来パイロットチャネルが１拡散コードを専有していたのに対し、本実施形態ではパイロットチャネルは実質１／４拡散コードのみ専有するのと等価となり、拡散コードの不足問題に対するパイロットチャネルの影響を低減できる。

【００１４】移動局は、パイロットチャネルとして使用しうる全ての拡散コードとタイムスロット＃の情報を自局のメモリに予め記憶している。移動局は電源立ち上げ時の在圏セル判定処理において、記憶している複数の拡散コードとタイムスロット＃について順次パイロットチャネルの受信レベル測定を行い、最も大きな受信レベルを有する拡散コードとタイムスロットでパイロットチャネルを送信しているセルを在圏セルとして判定する。待ち受け中および通信中の在圏セル判定処理においては、全ての隣接セルから送信されるパイロットチャネルの拡散コードおよびタイムスロット＃の情報が在圏セルの基地局から移動局に対して通知され、移動局は通知された複数の拡散コードとタイムスロット＃について順次パイロットチャネルの受信レベル測定を行い、最も大きな受信レベルを有する拡散コードとタイムスロットでパイロットチャネルを送信しているセルを在圏セルとして判定する。

【００１５】ここで他のパイロットチャネル配置方法として、パイロットチャネルを割り当てるタイムスロット＃を全セルで共通とする。これにより移動局の在圏セル判定処理に関し、タイムスロット番号に関する情報を移動局は拡散コード毎に記憶する必要がなくなる。また基地局から移動局に通知する隣接セルのパイロットチャネルに関する情報は拡散コードのみでよく、タイムスロット番号を不要にできる。

【００１６】基地局では、通信チャネルとして使用されるタイムスロットは、ＣＤＭＡ特有の問題である遠近問題を解決するために送信電力制御され、時間とともに送信電力は変化する。これに対しパイロットチャネルは移動局の在圏セル判定に用いるために一定送信電力でかつ通信チャネルより比較的大きい送信電力で送信される必要がある。よってパイロットチャネルとして割り当てた特定の拡散コードの特定のタイムスロットについては、常時一定送信電力で送信し、通信チャネルとして使用される他のタイムスロットの送信電力に対し比較的大きな送信電力で送信する。

【００１７】図２にパイロットチャネルを含む無線チャネルの送信電力の時間的な変動例を示す。図２に示すように、パイロットチャネルとして割り当てているタイムスロット＃１は常時一定の比較的大きな送信電力で送信され、他の通信チャネルとして使用しているタイムスロット＃２と＃３は送信電力制御され、時間とともに送信電力値は変化する。タイムスロット＃４は未使用であり送信されていない。

【００１８】従来はパイロットチャネルが１無線チャネルを専有していたため、常時一定送信電力でかつ通信チャネルより比較的大きい送信電力で送信されていた。これにより隣接セルへ大きな干渉を与えていた。これに対し本実施形態ではパイロットチャネルは全時間の１／４しか送信されないため、実質他セルへのパイロットチャネルによる干渉電力は従来の１／４となり、他セルに対する容量劣化の影響を低減できる。

【００１９】

【発明の効果】以上説明したように、本発明によれば、下り無線チャネルを複数のタイムスロットに時間分割して時間分割多重し、複数のタイムスロットの内の１つをパイロットチャネルとして割り当てることにより、従来パイロットチャネルが１拡散コードを専有していたのに対し、本発明では実質的に１拡散コードの時間多重数（１フレーム内のタイムスロット数）分の１のみ専有することとなるため、拡散コード不足に対するパイロットチャネルの影響を低減できる。

【００２０】また、本発明によれば、パイロットチャネルとして割り当てるタイムスロットの時間的な位置を、全セルで共通とすることにより、移動局の在圏セル判定処理に関し、タイムスロット番号に関する情報を移動局は拡散コード毎に記憶する必要がなくなる。また基地局から移動局に通知する隣接セルのパイロットチャネルに

関する情報は拡散コードのみでよく、タイムスロット番号を不要にできる。

【００２１】更に、本発明によれば、パイロットチャネルとして割り当てる前記タイムスロットのみ常時一定の送信電力で送信し、移動局との通信用の通信チャネルとして用いる他のタイムスロットは送信電力制御を行うことにより、従来パイロットチャネルが１無線チャネルを専有し、常時一定送信電力でかつ通信チャネルより比較的大きい送信電力で送信されていたのに対し、本発明ではパイロットチャネルは全時間の時間多重数（１フレーム内のタイムスロット数）分の１のみしか送信されないため、他セルへのパイロットチャネルによる干渉電力は従来の時間多重数分の１となり、他セルに対する容量劣化の影響を低減できる。

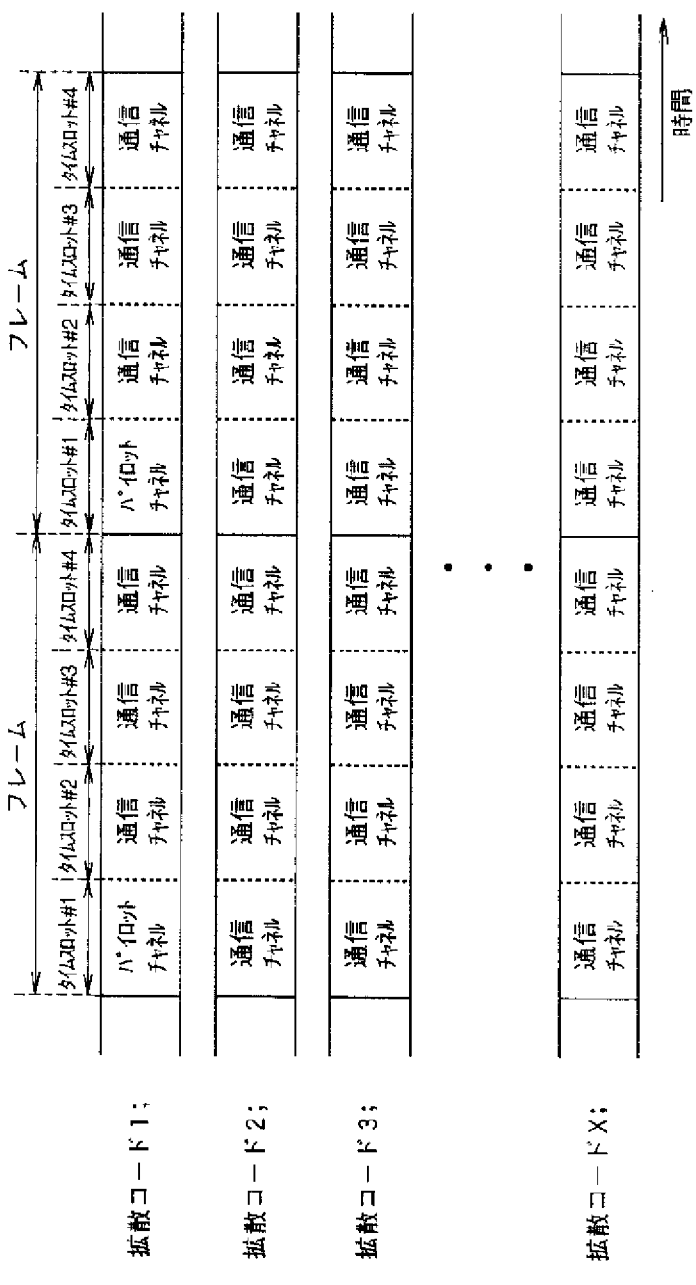
【図面の簡単な説明】

【図１】本発明の一実施形態に係るＣＤＭＡ移動通信システムにおけるパイロットチャネル配置および送信方法を実施するパイロットチャネルの配置方法の一例を示す図である。

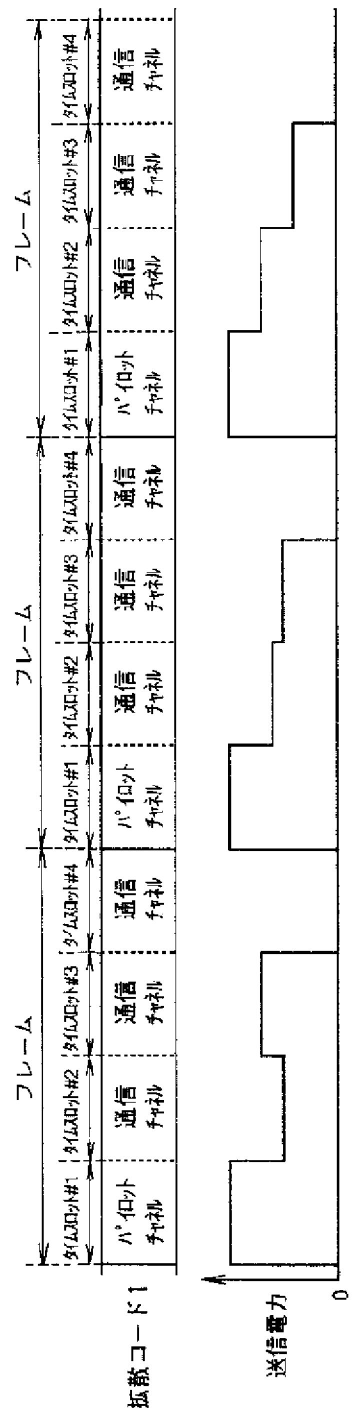
【図２】本発明の他の実施形態におけるパイロットチャネルの送信方法の一例を示す図である。

【図３】従来のパイロットチャネルの配置方法を示す図である。

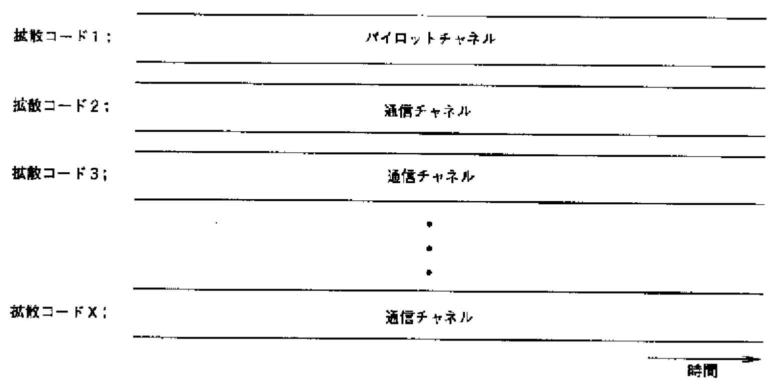
【図１】



【図２】


フレーム
タイムスロット#1
タイムスロット#2
タイムスロット#3
タイムスロット#4
パイロットチャネル
通信チャネル
拡散コード１
送信電力
0

【図３】



拡散コード１：
パイロットチャネル
拡散コード２：
通信チャネル
拡散コード３；
通信チャネル
拡散コードX：
通信チャネル
時間

JP9512156A

Bibliography

DWPI Title

Multiple carrier modulation and demodulation of digitally encoded data establishing safety intervals for control symbols in preamble of each frame larger than those for data symbols in data part of frame

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Application Number / Date

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DE4425713A / 1994-07-20 / DE
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WO1995EP2868A / 1995-07-20 / EP

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特表平9－512156

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(87)国際公開番号　ＷＯ９６／０２９８９
(87)国際公開日　平成８年(1996)２月１日
(31)優先権主張番号　Ｐ４４２５７１３．９
(32)優先日　1994年７月20日
(33)優先権主張国　ドイツ（ＤＥ）
(81)指定国　ＥＰ(ＡＴ，ＢＥ，ＣＨ，ＤＥ，ＤＫ，ＥＳ，ＦＲ，ＧＢ，ＧＲ，ＩＥ，ＩＴ，ＬＵ，ＭＣ，ＮＬ，ＰＴ，ＳＥ)，ＡＵ，ＣＡ，ＣＺ，ＦＩ，ＨＵ，ＪＰ，ＫＲ，ＰＬ，ＵＳ

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(74)代理人　弁理士　田中　浩　（外２名）

(54)【発明の名称】　持続波ネットワークにおける多重キャリア伝送

(57)【要約】
ディジタルコード化データでＯＦＤＭ変調を行うことによって所要の信号伝送容量が減少するのを防止するために、各ＯＦＤＭ伝送フレームのフレーム・ヘッダ中の制御シンボル保護期間を各ＯＦＤＭ伝送フレームの所要データ領域中のデータ・シンボル保護期間より長くする。データ・シンボル保護期間の長さは復調時に求められ、次いでデータ・シンボル検出タイミングがその保護期間の長さに応じて求められる。

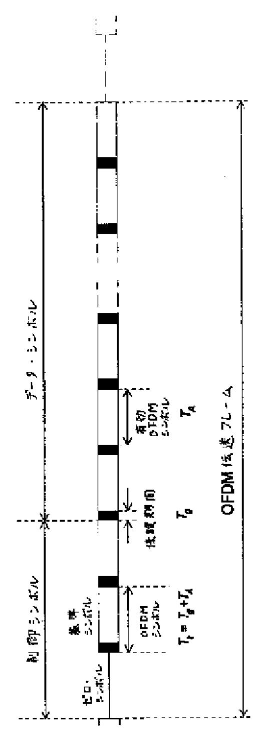


図１

【特許請求の範囲】

１．連続するＯＦＤＭ伝送フレームの形式に構成されたＯＦＤＭシンボルの時間シーケンスが生成され、上記ＯＦＤＭ伝送フレームの各々が、１つ以上の制御シンボルを有するフレーム・ヘッダと複数のデータ・シンボルを有する有効データ領域とからなり、連続する各制御シンボルまたは各データ・シンボルの間にそれぞれ保護期間を有する、ディジタル符号化データで多重キャリア変調を行う方法において、

上記各ＯＦＤＭ伝送フレームの上記フレーム・ヘッダにおける上記制御シンボルに対する保護期間を、上記各ＯＦＤＭ伝送フレームの上記有効データ領域における上記データ・シンボルに対する保護期間よりも長くなるように選択することを特徴とする、ディジタル符号化データで多重キャリア変調を行う方法。

２．上記各ＯＦＤＭ伝送フレームの上記有効データ領域における上記データ・シンボルに対する上記保護期間の時間的長さが可変であることを特徴とする、請求項１に記載のディジタル符号化データで多重キャリア変調する方法。

３．請求項１または２に記載された方法に従って変調された多重キャリアを復調して元のディジタル符号化データを復元する方法において、受信した上記各ＯＦＤＭ伝送フレームの上記有効データ領域における上記データ・シンボルに対する保護期間の長さを求め、上記データ・シンボルの各検出タイミングを、検出した上記保護期間の長さの関数として決定することを特徴とする、多重キャリア復調して元のディジタル符号化データを復元する方法。

【発明の詳細な説明】

持続波ネットワークにおける多重キャリア伝送

発明の詳細な説明

発明の属する技術分野

本発明は、請求の範囲の請求項１に従うディジタル符号化データで多重キャリア変調する方法と、そのようにして多重キャリア変調された信号を、請求項３に従って復調して元のディジタル符号化データを復元する方法に関する。この前提技術の変復調方法は、フランス特許出願公開明細書ＦＲ－Ａ－2,639,495号により周知である。

発明の背景

フランス特許出願公開明細書ＦＲ－Ａ－2,639,495号により公知の多重（多、多数）キャリア変調（直交周波数分割多重変調(orthogonal frequency division multiplex modulation)、ＯＦＤＭ変調と略称する）において、ＯＦＤＭシンボル（符号）の時間シーケンスは、ディジタル符号化データで変調した多数のキャリア（搬送波）をフーリエ解析することによって生成される。その各ＯＦＤＭシンボルは連続するＯＦＤＭ伝送フレームの形式に配置構成され、その各ＯＦＤＭ伝送フレーム間は例えばゼロ（０）シンボルまたは空のシンボル（図１）等によって互いに隔てられ（分離し）ている。各ＯＦＤＭ伝送フレームは、１つまたは数個の制御シンボルを有するフレーム・ヘッダと、その後に多数のデータ・シンボルを有する有効データ領域（有用データ領域）とからなる。ＯＦＤＭ復号器においては、制御シンボルを使用して、各受信ＯＦＤＭ伝送フレームの開始点と各ＯＦＤＭシンボルを適正なタイミング（時間）で検出するとともに、レベル（または振幅）と位相に応じて正確な変調キャリア周波数を復元する。変調器側では、連続するＯＦＤＭシンボル（制御シンボルおよびデータ・シンボル）の各シンボル相互間に保護（ガード）期間が挿入される。その保護期間が存在することによって、マルチパス（多重通路）伝播に起因して復調器側で生じる連続ＯＦＤＭシンボル間のクロストークまたは干渉を防止することができる。この公知文献では、各ＯＦＤＭ伝送フレームにおけるフレーム・ヘッダおよび有効データ領域における全てのＯＦＤＭシンボルに対する保護期間の時間的長さは互いに同じであ

る。

一方、特に持続波（同一周波）ネットワーク(Gleichwellennetzen)における受信位置において各遅延時間の間に大きな差が生じる場合は、その保護期間は比較的長い持続時間を有するように設計して、連続する各ＯＦＤＭシンボル相互間のクロストークを高い信頼性で防止するようにしなければならない。しかし、そのように長い持続時間の保護期間を設けると、その結果として、有効信号の伝送容量または伝送効率が減少する。そのような状況を改善するための１つの選択肢として、保護期間の長さを長くし、また有効信号期間の時間長を同じ程度長くすればよい。しかし、そのようにすると、復調器側の費用（コスト）が相当高くなる。即ち、要求される、ＯＦＤＭデータ・シンボルの検出（走査）精度、検出値の記憶容量、およびその検出値から得られる時間信号の周波数解析の計算費用（コスト）が、それぞれ不相応に過大に高くなる。従って、ＯＦＤＭ変調システムのための保護期間は、ネットワーク計画設計の観点から伝送容量、受信機の費用（コスト）および周波数効率に対する欠点を容認するように妥協して選択しなければならない。

これに対して、本発明の目的は、大多数のアプリケーション（適用例）において有効信号の伝送容量を減少させないようにし、同時に、広域に拡がる持続波ネットワークの場合をも考慮した伝送方式を実現することである。

<u>発明の概要</u>

この目的は、請求の範囲の独立請求項１および３の発明の特徴によって解決される。

本発明は、必ずしも全ての放送サービスまたは同報サービスに対して長い保護期間を設ける必要はないという認識に基づいている。例えば、ローカル・ネットワーク構成と地域的(regional)ネットワーク構成と全国的(national)ネットワーク構成との間では、必要な保護期間の長さは相違する。最も重要な点は、広い範囲の地域に持続波動作で信号を供給することである。本発明の思想は、このような認識から始まったもので、各ＯＦＤＭ伝送フレームのフレーム・ヘッダに対する保護期間だけを、考え得る最悪の条件のアプリケーションを想定して設計し、

各ＯＦＤＭ伝送フレームの有効データ領域のＯＦＤＭシンボルに対する保護期間は、考慮対象のネットワークにおいて生じる実際の遅延時間差に応じて設計する

ことである。

図面の簡単な説明

図１は、従来技術によるＯＦＤＭ伝送フレームの概略設計を示す。

図２は、本発明に従うＯＦＤＭ伝送フレームの概略設計の３つの例を示す。

図３は、図２に従ってＯＦＤＭ伝送フレームを生成するＯＦＤＭ変調器のブロック図を示す。

図４は、図２に従うＯＦＤＭ伝送フレームの変調信号を復調するＯＦＤＭ復調器のブロック図を示す。

発明の実施態様の詳細な説明

次に、本発明を図面を参照して詳細に説明する。

図３は、送信側のＯＦＤＭ変調の基本的な特徴を示している。最小の伝送単位を表すＯＦＤＭシンボルを形成するために、ディジタル符号化データのブロック３１０を考える。そのデータを用いて、周波数領域３２０における複数の直交キャリアが変調される。この処理において、標準的ディジタル変調技術（例、ＱＰＳＫ、ＱＡＭ、等）を適用して各キャリアを変調する。逆フーリエ変換３３０を用いて、持続時間T_Aの周期（期間）の時間信号を合成する。この周期的時間信号は、データ・ブロック３１０の全体の情報を含んでおり、中間的に記憶された検出値３４０の形態で利用可能となるものであって、アナログ信号に変換され、シンボル持続時間T_sで送信される。ここで、シンボル持続時間T_sは、サイクル持続時間（周期）によって決まる最小限の有効シンボル時間T_Aより長くなるように選択される。送信信号に対する付加時間または送信信号の延長時間は、持続時間$T_g - T_s - T_A$を有する保護期間(guard interval)として表される。

本発明によれば、制御シンボル(Steuerungssymbole)に対する保護期間$T_{g\ strg}$（$T_{g\ cont}$）は、データ・シンボルに対する保護期間$T_{g\ data}$より長い長さに選択される。この選択は、ＯＦＤＭ伝送フレームにおけるＯＦＤＭシンボルの位置の関数の形で制御される切換え手段３６０によって実行される。ここで、保護

期間$T_{g\ strg}$の長さは、広域の持続波ネットワークにおいて予想（想定）される臨界条件においても連続する制御シンボル間のクロストークが回避（防止）できるように選択される。保護期間$T_{g\ data}$は、相異なる保護期間の集合（群）３７

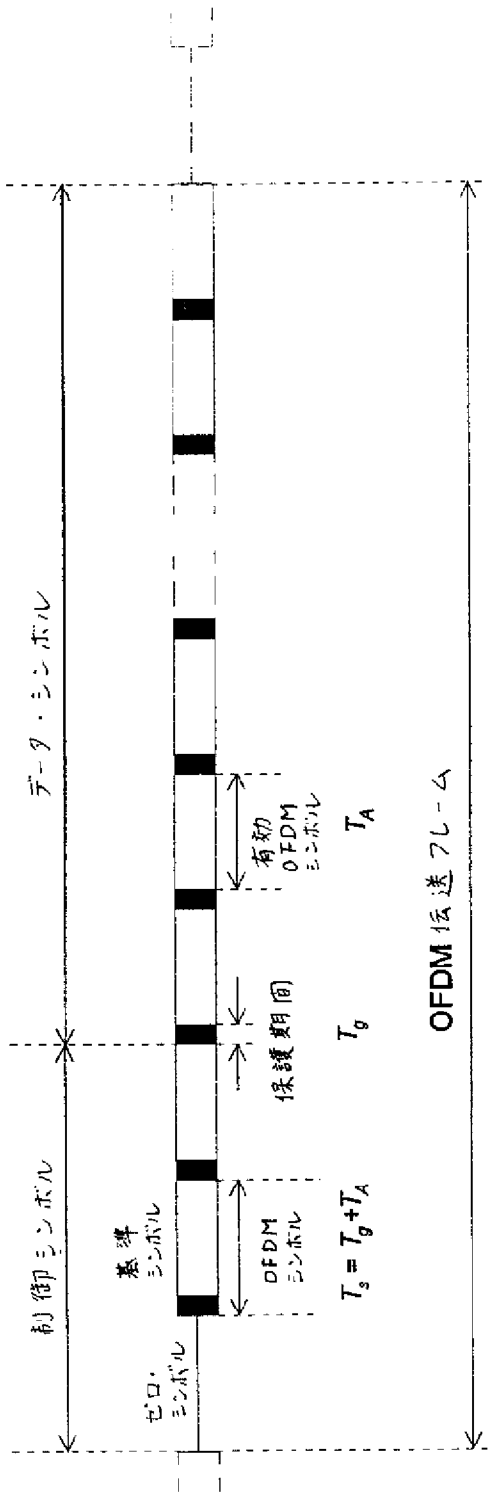
０の中から選択することができるものであり、個々の事例において実現した送信機のネットワーク構成によって実際に各遅延時間に差が生じたときにも、連続するデータ・シンボル間に実質的にクロストークが生じないような長さに設定される。データ・シンボルの保護期間$T_{g\ data}$に対して選択した持続時間の長さは、制御シンボルの信号を用いて受信側に伝えられる。このようにして、例えば図２のａ）～ｃ）に示すような相異なる保護期間を有するＯＦＤＭ伝送フレームが生成される。

受信側では、最初にＯＦＤＭ伝送フレームに対して概略の粗い同期が取られる。受信した時間信号４００は、最初に伝送されたＯＦＤＭシンボルのシンボル持続時間T_sにおいて、検査（探査）されて最初にサイクル持続時間T_Aが検出され（４２０）、周波数領域へのフーリエ変換（４３０）によって解析される。伝送されたデータ４５０は、個々のキャリアを復調して復元される。一方、精細な時間同期を取って同期を微調整するために、通常、最初（第１の）ＯＦＤＭシンボルが基準シンボルとして用いられて、伝送チャンネルのインパルス応答の計算が可能となり、かつキャリア周波数の振幅および位相が復元できる。伝送チャンネルのインパルス応答を用いることによって、長い時間となるように選択された保護期間$T_{g\ strg}$の持続時間の範囲内でマルチパス・チャンネルおよび持続波ネットワークにおいて生じる全てのエコーの振幅と遅延時間を知ることができる。

後続の各ＯＦＤＭシンボルを検査してサイクル持続時間T_Aを検出するタイミングは、伝送フレームにおけるＯＦＤＭシンボルの位置の関数として、それぞれの保護期間の持続時間分だけ遅延される。制御４９０において、各制御シンボルおよび各データ・シンボルに対する保護期間の持続時間は、本発明に従って予め設定されている。そのデータ・シンボルに対する保護期間$T_{g\ data}$の持続時間は、先に、信号（制御シンボル）で伝えられて、取り得る相異なる保護期間４９５の集合（群）の中から選択される。

本発明による方法の利点は、比較的少ないハードウェア費用で保護期間の持続時間を柔軟に設定することができ、それによってＯＦＤＭシステムの実現およびネットワーク計画の観点からＯＦＤＭシステムを最適化することが可能になることである。また、本発明による方法の別の利点は、データ・シンボルに対して予め設定した保護期間を越えるエコー遅延時間の差が生じたときに明らかになる。制御シンボルに対して充分に長い保護期間が与えられている限り、さらに、基準シンボルを用いて求められる伝送チャンネルのインパルス応答を用いた通常の方法により、データ・シンボルに対してエコー等化を行うことも可能となる。

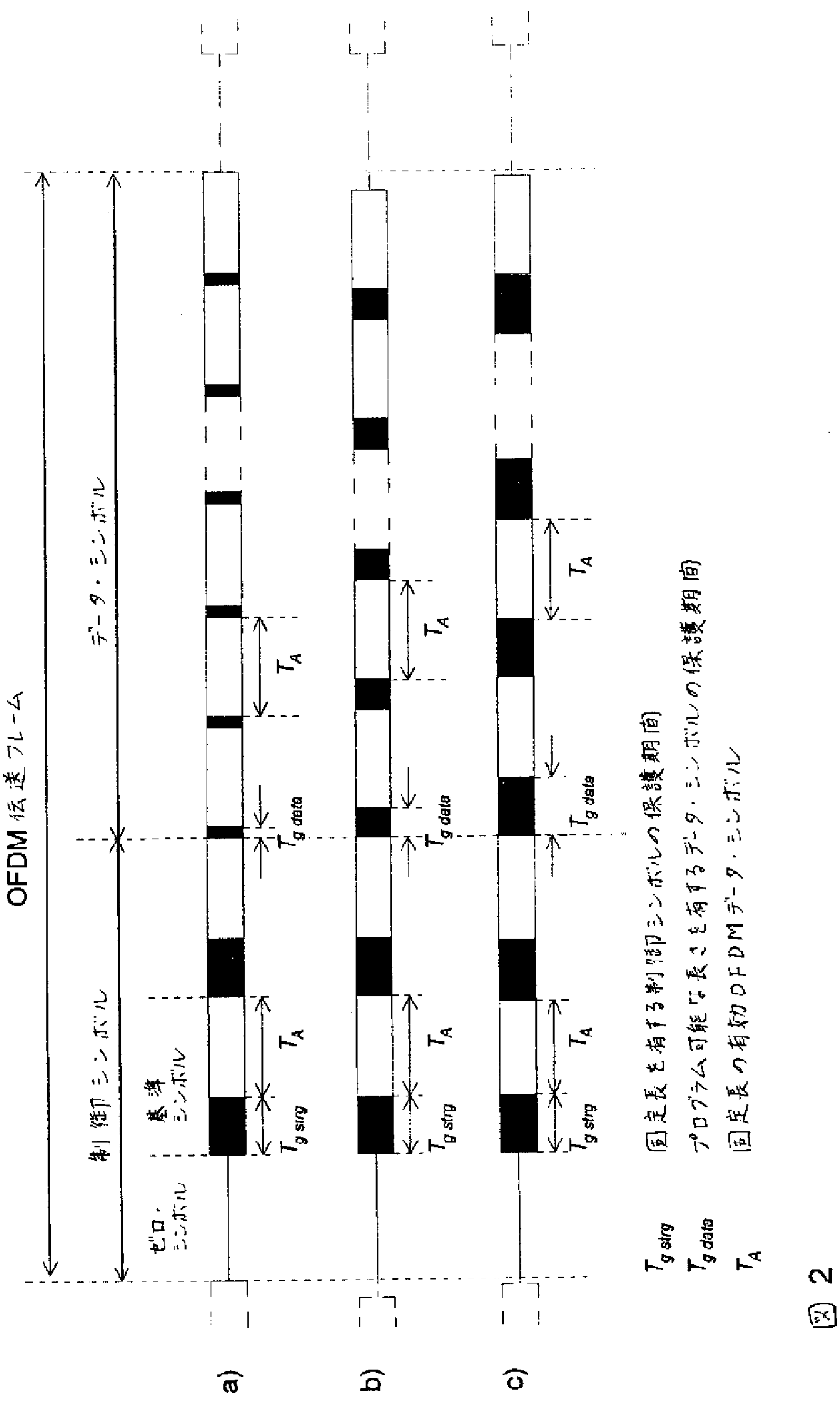
【図１】



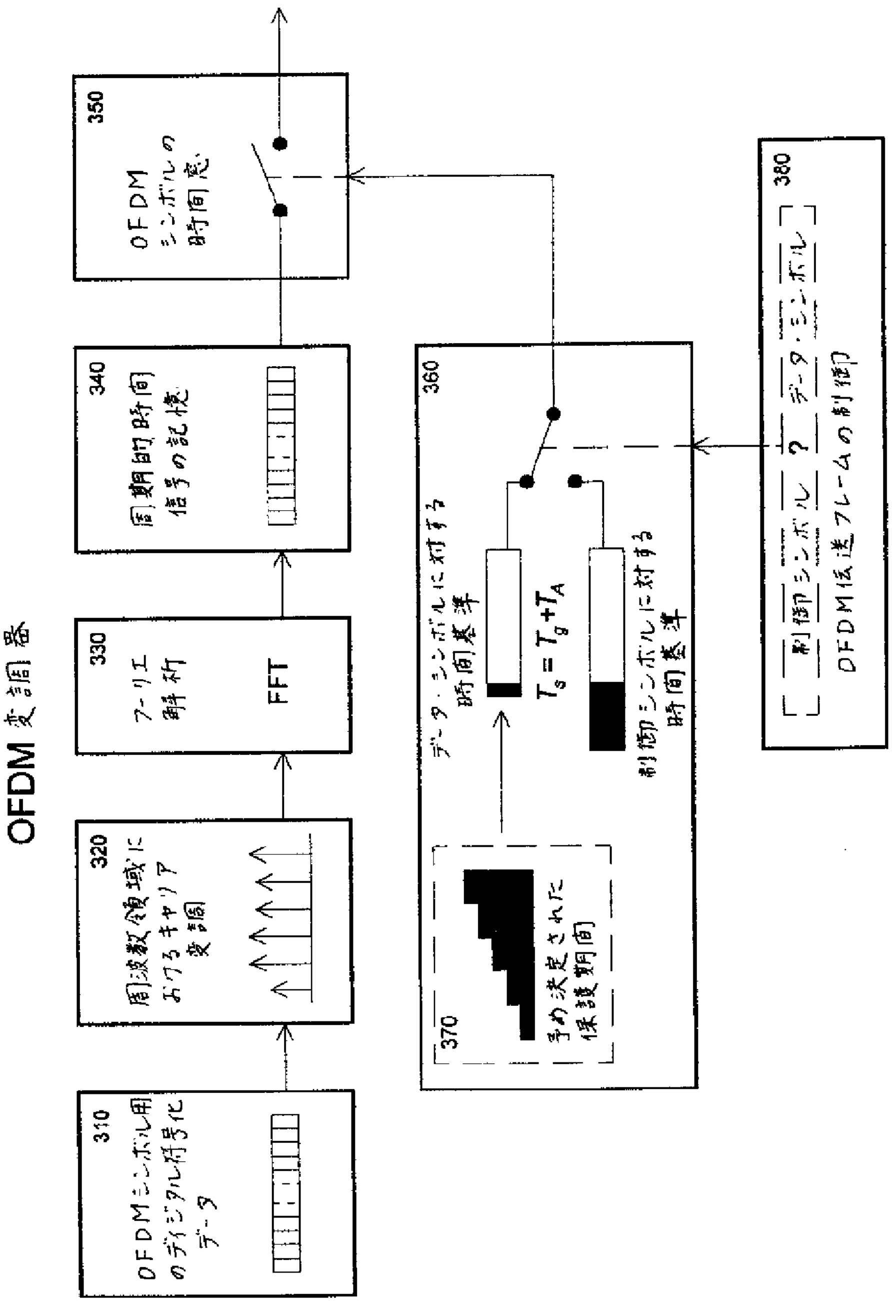
制御シンボル
データ・シンボル
ゼロ・シンボル
基準シンボル
OFDMシンボル
保護期間
有効OFDMシンボル
$T_s = T_g + T_A$
T_g
T_A
OFDM伝送フレーム

図１

【図2】


OFDM伝送フレーム
データ・シンボル
制御シンボル
ゼロ・シンボル
基準シンボル
$T_{g\,strg}$
$T_{g\,data}$
T_A
a)
b)
c)
$T_{g\,strg}$ 固定長を有する制御シンボルの保護期間
$T_{g\,data}$ プログラム可能な長さを有するデータ・シンボルの保護期間
T_A 固定長の有効OFDMデータ・シンボル
図2

【図3】



OFDM変調器
310
OFDMシンボル用のディジタル符号化データ
320
周波数領域におけるキャリア変調
330
フーリエ解析
FFT
340
同期時間信号の記憶
350
OFDMシンボルの時間窓
360
データ・シンボルに対する時間基準
$T_s = T_g + T_A$
制御シンボルに対する時間基準
370
予め決定された保護期間
380
制御シンボル ? データ・シンボル
OFDM伝送フレームの制御

図3

【図４】

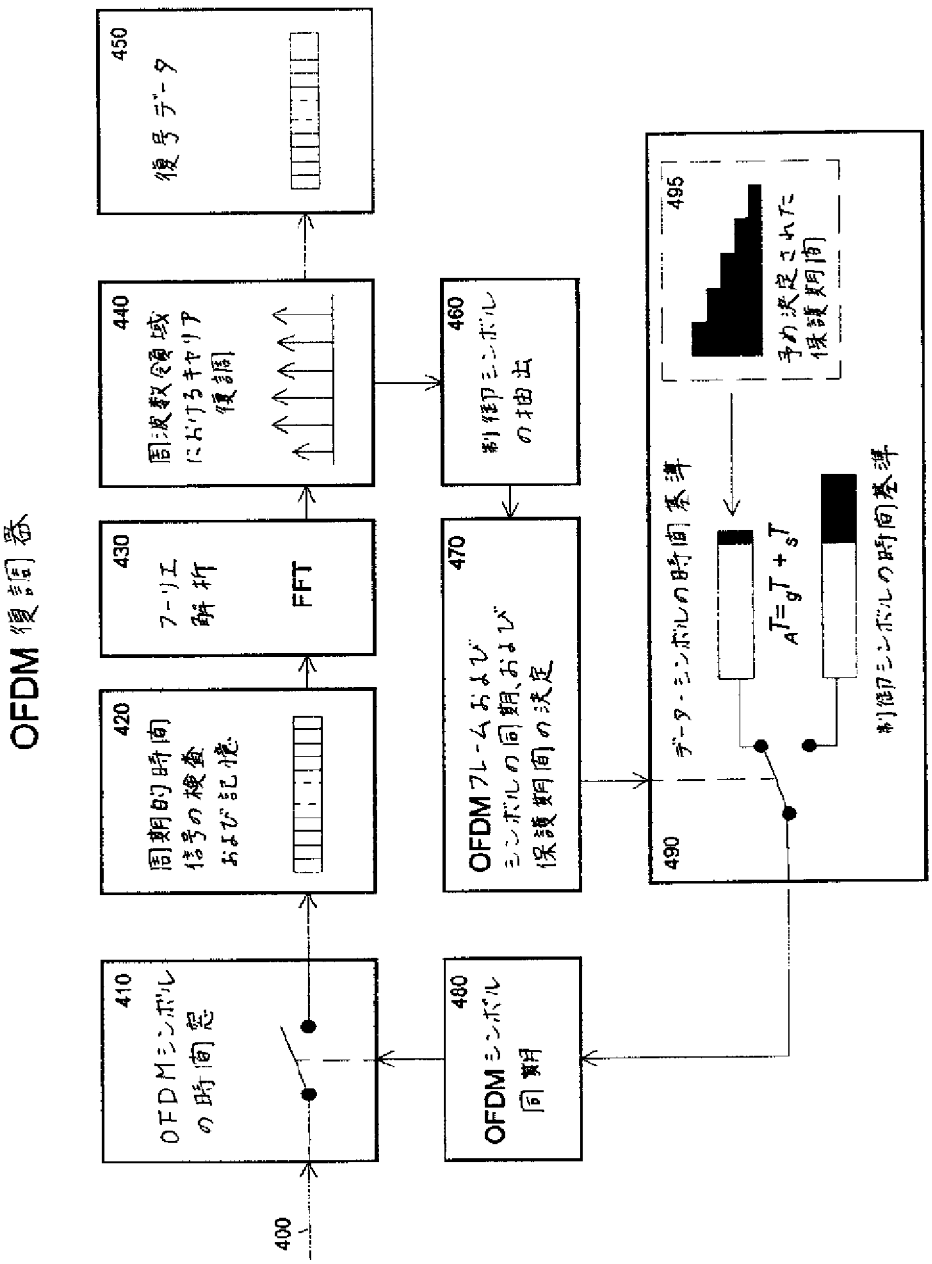

OFDM復調器
400
410
OFDMシンボルの時間窓
420
同期的時間信号の検査および記憶
430
フーリエ解析
FFT
440
周波数領域におけるキャリア復調
450
復号データ
460
制御シンボルの抽出
470
OFDMフレームおよびシンボルの同期、および保護期間の決定
480
OFDMシンボル同期
490
データ・シンボルの時間基準
$T_A = T_g + T_s$
制御シンボルの時間基準
495
予め決定された保護期間

図４

【国際調査報告】

INTERNATIONAL SEARCH REPORT

Interm al Application No
PCT/EP 95/02868

A. CLASSIFICATION OF SUBJECT MATTER
IPC 6 H04L5/06 H04H3/00

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 6 H04L H04H

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	RUNDFUNKTECHNISCHE MITTEILUNGEN, vol. 38, no. 1, January 1994 NORDERSTEDT DE, pages 14-23, BRUGGER 'DAB - Gleichwellennetze bei 1,5GHz' see page 14, right column, paragraph 2 - page 15, left column, paragraph 2 see table 1 --- -/--	1-3

[X] Further documents are listed in the continuation of box C.

[ ] Patent family members are listed in annex.

* Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier document but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
3 November 1995	20. 11. 95

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax (+31-70) 340-3016	Scriven, P

Form PCT/ISA/210 (second sheet) (July 1992)

INTERNATIONAL SEARCH REPORT

International Application No
PCT/EP 95/02868

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	SIGNAL PROCESSING IMAGE COMMUNICATION., vol. 5, no. 5/6, December 1993 AMSTERDAM NL, pages 379-403, TOURTIER ET AL. 'Multicarrier modem for digital HDTV terrestrial broadcasting' see figures 6-8,11 see page 385, paragraph 2 - page 387, paragraph 1 ---	1-3
A	IEEE TRANSACTIONS ON CONSUMER ELECTRONICS, vol. 35, no. 3, August 1989 NEW YORK, US, pages 493-503, LE FLOCH ET AL 'DIGITAL SOUND BROADCASTING TO MOBILE RECEIVERS' see figure 3 see page 496, left column, paragraph 1 see page 501, right column, paragraph 5 -----	1-3

Form PCT/ISA/210 (continuation of second sheet) (July 1992)

JP10327126A CDMA RECEIVER

Bibliography

DWPI Title

Code division multiple access receiver used in mobile communication subtracts corresponding cancellation signals produced by different ones of other path demodulators from CDMA signal associated with that subtractor

Original Title

CDMA RECEIVER

Assignee/Applicant

Standardized: LUCENT TECHNOLOGIES INC
Original: LUCENT TECHNOL INC

Inventor

HUANG HOWARD C ; CHIH-LIN I ; TEN BRINK STEPHAN ; VANNUCCI GIOVANNI

Publication Date (Kind Code)

1998-12-08 (A)

Application Number / Date

JP1998116412A / 1998-04-27

Priority Number / Date / Country

US1997841316A / 1997-04-30 / US
JP1998116412A / 1998-04-27 / JP

Abstract

PROBLEM TO BE SOLVED: To improve performance by preventing interference to a multipath caused by a pilot signal by removing the pilot signal of multipath component, which causes the interference of base band received signal, by reconstituting it as prescribed and adding/subtracting that signal later.

SOLUTION: A signal r0 of path 0 and a signal r1 of path 1 in received decoding signals r(n) are respectively processed by an on-time selector circuit OTS and inputted to RAKE fingers 603 and 604 later. In this case, pilot reconstitution circuits 606 and 607 are reconstituting pilot signals through predicted attenuation, phase and path delay are respectively added to the fingers 603 and 604, the pilot signal from the path 0 is reconstituted, subtracted from the signal of path 1 by an adder 609 and removed. Similarly, the pilot signal reconstituted from the path 1 is subtracted from the signal of path 0 by an adder circuit 608 and removed and afterwards, the signals are respectively exactly demodulated by fingers 0 and 1 and bit-determined or processed by a viterbi decoder 605.

(19)日本国特許庁（ＪＰ）　(12) 公開特許公報（Ａ）　(11)特許出願公開番号

特開平10－327126

(43)公開日　平成10年(1998)12月８日

(51)Int.Cl.^{6}	識別記号	ＦＩ	
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Ｈ０４Ｂ　7/26		Ｈ０４Ｌ　7/00	Ｃ
Ｈ０４Ｌ　7/00		Ｈ０４Ｂ　7/26	Ｄ
			Ｐ

審査請求　未請求　請求項の数24　ＯＬ　（全 25 頁）

(21)出願番号　特願平10－116412

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(31)優先権主張番号　０８／８４１３１６

(32)優先日　1997年４月30日

(33)優先権主張国　米国（ＵＳ）

(71)出願人　596077259
ルーセント　テクノロジーズ　インコーポレイテッド
Ｌｕｃｅｎｔ　Ｔｅｃｈｎｏｌｏｇｉｅｓ　Ｉｎｃ．
アメリカ合衆国　07974　ニュージャージー、マレーヒル、マウンテン　アベニュー　600－700

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(74)代理人　弁理士　三俣　弘文

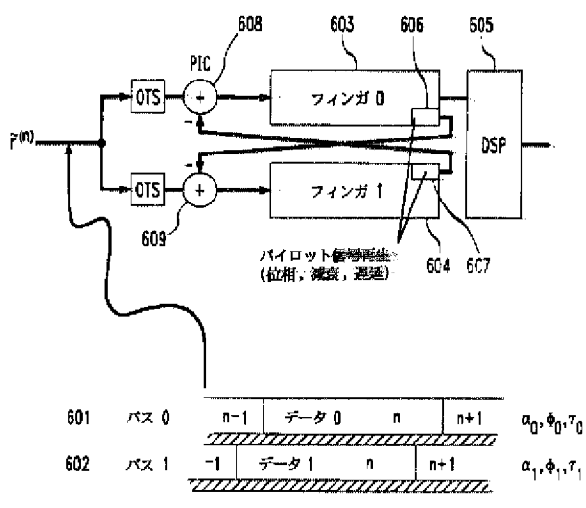
最終頁に続く

(54)【発明の名称】　ＣＤＭＡ受信機

(57)【要約】

【課題】パイロット信号干渉除去技術を用いたコヒーレントＭＣ－ＣＤＭＡ受信機を提供すること。

【解決手段】　本発明は、ユーザデータチャネルと、Ｌ個のパスを介しての個々のパイロットチャネルとを含むコヒーレントＣＤＭＡ信号を受信し復調するＣＤＭＡ受信機において、前記所望のデータチャネルは、あるパスのパイロットチャネルとは直交し、（Ａ）１個のパスを介して受信したＣＤＭＡ信号からデータチャネルとパイロットチャネルを見積り減算手段が使用する（Ｌ－１）個の除去信号を生成するＬ個のパス復調器と、（Ｂ）自己の減算手段に関連しない他の（Ｌ－１）個のパス復調器のうちの個々の復調器により生成された（Ｌ－１）個の除去信号を、減算手段に関連するＣＤＭＡ信号から減算するＬ個の減算手段とを有することを特徴とする。



【特許請求の範囲】

【請求項１】　少なくとも１個のユーザデータチャネルと、複数のＬ個（Ｌは２以上の整数）のパスを介しての個々のパイロットチャネルとを含むコヒーレントＣＤＭＡ信号を受信し復調するＣＤＭＡ受信機において、

前記所望のデータチャネルは、あるパスのパイロットチャネルとは直交し、

（Ａ）　Ｌ個のパスのうちの１個のパスを介して受信したＣＤＭＡ信号からデータチャネルとパイロットチャネルを見積り、減算手段が使用する（Ｌ－１）個の除去信号を生成するＬ個のパス復調器と、

（Ｂ）　関連減算手段に関連しない他の（Ｌ－１）個のパス復調器のうちの個々の復調器により生成された（Ｌ－１）個の除去信号を、減算手段に関連するＣＤＭＡ信号から減算するＬ個の減算手段とを有することを特徴とするＣＤＭＡ受信機。

【請求項２】　前記（Ｌ－１）個の除去信号は、再構成されたパイロット信号であり、

前記各（Ｂ）減算手段は、前記再構成されたパイロット信号を復調器へ入力される信号から減算するために、前記関連復調器の前に配置されることを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項３】　前記（Ｌ－１）個の除去信号は、一対の相関処理で再構成されたパイロット信号であり、

前記（Ｂ）減算手段は、一対の相関処理で再構成されたパイロット信号をそのデータ／パイロットアキュームレータから出力された信号から減算するために、その復調器のデータ／パイロットアキュームレータの後に配置された一対の減算器であることを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項４】　前記除去用のパイロット信号は、１個のシンボルに対し実行された第１チャネル見積により得られた最新のチャネル見積を有するチャネル見積アルゴリズムを用いて再構成され、

前記復調器の入力は、１個のシンボルの間、来入するチップレート信号をバッファリングすることにより得られることを特徴とする請求項２記載のＣＤＭＡ受信機。

【請求項５】　前記除去用のパイロット信号は、前のシンボル間隔復調から得られた最新のチャネル見積を有するチャネル見積アルゴリズムを用いて再構成され、

これらのチャネル見積は、パイロット信号の再構成と前のシンボル間隔の復調用に用いられることを特徴とする請求項２記載のＣＤＭＡ受信機。

【請求項６】　第１チャネル見積の前の減算手段は、１個のシンボルに亘って、前のシンボル間隔復調から得られた最新のチャネル見積を有するチャネル見積アルゴリズムを用いて再構成されたパイロット信号を除去することを特徴とする請求項４記載のＣＤＭＡ受信機。

【請求項７】　前記一対の相関処理で再構成された除去用のパイロット信号は、利用可能な最新のチャネル見積を有するチャネル見積アルゴリズムを用いて再構成されることを特徴とする請求項３記載のＣＤＭＡ受信機。

【請求項８】　チャネル見積アルゴリズムの前に減算手段を有し、この減算手段は一対の相関処理されたパイロット信号を再構成するために用いられる出力を有し、パイロットアキュームレータ信号上の除去用に用いられる一対の相関処理されたパイロット信号の成分の第１中間結果を除去することを特徴とする請求項７記載のＣＤＭＡ受信機。

【請求項９】　（Ｃ）　他の（Ｌ－１）個の復調器の遅延時間に関連するパイロット除去信号のパルス形状を再構成する再構成用ローパスフィルタ（ＲＬＰ）をさらに有することを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項１０】　前記（Ｃ）再構成用ローパスフィルタは、有限インパルス応答（ＦＩＲ）フィルタを用いて実現されることを特徴とする請求項９記載のＣＤＭＡ受信機。

【請求項１１】　前記有限インパルス応答フィルタは、ルックアップテーブルを用いて実現されることを特徴とする請求項１０記載のＣＤＭＡ受信機。

【請求項１２】　前記再構成用ローパスフィルタは、係数用のルックアップテーブルを用いて実現されることを特徴とする請求項９記載のＣＤＭＡ受信機。

【請求項１３】　（Ｄ）　指定されたマルチパス成分の信号パワーに従って、（Ｌ－１）個の除去信号の生成を入り切りする各復調器内のスイッチ手段をさらに有することを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項１４】　前記パイロットチャネルは、少なくとも１個のユーザ信号チャネルに直交することを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項１５】　前記パイロットチャネルは、あるパスの所望のユーザ信号チャネルに非直交であり、

各復調器は、復調される前にマルチパス成分の非直交パイロット信号を除去するために（Ｌ－１）個の除去信号と付属の除去信号を生成し、

前記非直交パイロット信号の除去は、各Ｌ個の減算手段内の余分の付属的減算を用いて行われることを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項１６】　前記複数の信号チャネルは、ウォルシュ符号を用いて符号化されることを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項１７】　少なくとも一人のユーザは、複数の信号チャネルを使用することを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項１８】　前記コヒーレントＣＤＭＡ信号は、少なくともＱ信号チャネルとＩ信号チャネルを含むことを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項１９】　ＣＤＭＡシステムのユーザ局の一部は、少なくとも１つの基地局と複数のユーザ局とを含む

ことを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項２０】　ＣＤＭＡシステムの基地局の一部は、少なくとも１つの基地局と複数のユーザ局とを含むことを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項２１】　減算手段の出力をチャネル重み付けする手段と、

前記重み付けされた出力を結合する手段をさらに有することを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項２２】　少なくとも１個のユーザデータチャネルと、複数のＬ個（Ｌは２以上の整数）のパスを介しての個別のパイロットチャネルとを含むコヒーレントＣＤＭＡ信号を受信し、復調するＣＤＭＡ受信機の動作方法において、

前記所望のデータチャネルは、あるパスのパイロットチャネルとは直交し、

（Ａ）　Ｌ個のパス復調器の各々において、Ｌ個のパスのうちの１個のパスを介して受信したＣＤＭＡ信号からデータチャネルとパイロットチャネルを見積り、特定の減算手段により使用される（Ｌ－１）個の除去信号を生成するステップと、

（Ｂ）　Ｌ個の減算手段の各々において、自己の減算手段に関連しない他の（Ｌ－１）個のパス復調器のうちの別の１個の復調器により生成された（Ｌ－１）個の除去信号を減算手段に関連するＣＤＭＡ信号から減算するステップとからなることを特徴とするＣＤＭＡ受信機の動作方法。

【請求項２３】　（Ｌ－１）個の除去信号の１つまたは複数の減算を制御する手段、

をさらに有し、

前記（Ｌ－１）個の除去信号の各々の減算は、関連データチャネル信号で受信されたパイロット信号とその変動分に基づいて決定されたしきい値レベルの関数として制御されることを特徴とする請求項１記載のＣＤＭＡ受信機。

【請求項２４】　生成され減算されるべき除去信号の組は、式（６）で与えられることを特徴とする請求項２３記載のＣＤＭＡ受信機。

【発明の詳細な説明】

【０００１】

【発明の属する技術分野】本発明は、符号分割多重（ＣＤＭＡ）受信機に関し、パイロット干渉除去技術を用いたコヒーレントＭＣ－ＣＤＭＡ受信機に関する。

【０００２】

【従来の技術】符号分割多重アクセス（ＣＤＭＡ）は、ワイアレス通信システムの最も有望なシステムになりつつある。ＣＤＭＡユーザは、異なる符号シーケンスにより他のユーザから区別されている。ＣＤＭＡ信号がワイドバンドである特徴により、この受信機は、ＲＡＫＥ受信機を用いて内蔵する時間ダイバスティを用いることによりフェージングに耐え得るようになっている。

【０００３】ＲＡＫＥ受信機をコヒーレントに実現するためには、パイロット信号を用いてコヒーレント検出に必要なチャネルの振幅と位相の予測値を得ている。ＩＳ－９５仕様のＣＤＭＡシステムの場合においては、このパイロット信号はユーザの拡散符号に対し直交しており、その結果マルチパスの分散がないという稀な場合においては、パイロット信号は、所望のユーザに対するマッチドフィルタの出力点で干渉を引き起こすことはない。

【０００４】しかし、マルチパス分散が存在する場合には、所望の信号に対し直交していない様々な種類のマルチパス成分に起因して、マッチドフィルタの出力点で不要な干渉が存在する。具体的に説明すると、所望のトラフィックチャネルのあるマルチパス成分にとっては、そのマッチドフィルタの出力は、他のマルチパス成分と、他のチャネルの他のマルチパス成分と、パイロット信号とに起因する不要な寄与分を有することになる。

【０００５】パイロット信号は、ダウンリンク信号のパワーの約２０％であるので、そのマルチパス成分は、活性トラフィックチャネルの全数が多い場合には、近遠効果（near-far effect）により、特に所望のユーザのビット決定に対し損傷を与えることがある。従来のＲＡＫＥ受信機は、チャネル間のマルチパス干渉に対し対策を講じていないため、その結果性能が劣化することになる。

【０００６】

【発明が解決しようとする課題】したがって本発明の目的は、パイロット信号干渉除去技術を用いたコヒーレントＭＣ－ＣＤＭＡ受信機を提供することである。

【０００７】

【課題を解決するための手段】本発明のＣＤＭＡ受信機は、受信信号からパイロット信号を除去する。このパイロット信号は、そのマルチパスパラメータ（振幅，位相シフトと遅延）と、そのシグネチャーシーケンス（signature sequence）によって規定されている。この情報はユーザの受信機端末（即ち、ハンドセット）に知られているので、ベースバンドの受信信号の干渉を来すマルチパス成分のパイロット信号が、検出され、所望のマルチパス成分を復調する前に取り除かれる。

【０００８】特に本発明のＣＤＭＡ受信機は、複数のＬ本（Ｌは２以上の整数）のパスを介して受信した個々のパイロットチャネルと、少なくとも１つのユーザデータチャネルを含むコヒーレントＣＤＭＡ信号を受信し復調する、そしてこの所望のデータチャネルは、あるパスではパイロットチャネルに対し直交している。

【０００９】ＣＤＭＡ受信機は、Ｌ個のパス復調器を有し、各復調器はＬ本のパスの内１本のパスを介して受信したＣＤＭＡ信号からデータチャネルとパイロットチャネルを予測し、Ｌ個の減算手段（subtractor means）の特定の１つにより使用される（Ｌ－１）個の除去信号

（cancellation signals）を生成する。各Ｌ個の減算手段は、その減算手段に関連するＣＤＭＡ信号から（Ｌ－１）個の除去信号（他の（Ｌ－１）個のパス復調器の別の復調器により生成される）を減算する。

【００１０】前復調型（pre-demodulation）の実施例においては、この（Ｌ－１）個の除去信号は、再構成されたパイロット信号であり、そして各減算手段は、その関連復調器の前に配置され、再構成パイロット信号をその復調器に入力される信号から減算する。

【００１１】後蓄積型（post-accumulation）の実施例においては、各（Ｌ－１）個の除去信号は、相関器で処理され再構成された一対のパイロット信号であり、各減算手段は一対の減算器であり、これらはその復調器のパイロットとデータのアキュムレータ（蓄積器）の後ろに配置され、一対の相関器処理された再構成パイロット信号をそのデータとパイロットアキュムレータからの出力信号から減算する。

【００１２】本発明の他の実施例においては、パイロット信号の除去は、所定レベルを超えて検出されたパス信号レベルに応じてオン、オフに切り換えられる。

【００１３】

【発明の実施の形態】図１にはＩＳ－９５用に与えられた値を有する同期パイロット符号補助のＣＤＭＡ通信リンク用の送信器が示されている。この実施例においては、レートR_b（ビットレート，シンボルレート）でユーザｊとｋからのデータ信号が符号拡散器１０１ｊと１０１ｋに接続されている。ここでは長さｇ（ｇ＝６４）の異なる直交Ｗａｌｓｈ符号が個々のユーザ用の拡散シーケントスとして用いられている。符号拡散器１０１ｊと１０１ｋの出力は、加算器１０２内でパイロット信号と結合されて、Ｗａｌｓｈ－拡散ベースバンド信号を形成する。最大（ｇ－１）個のデータチャネル（制御チャネルを含む）が一度に利用できる（チャネルの１つはパイロット信号により占有されている）。

【００１４】このＷａｌｓｈ－拡散ベースバンド信号は、例えばレートR_cで符号化器１０４－１０５内で別のＰＮ符号拡散シーケンス（ショートコードあるいはパイロット符号シーケンスとも称する）と乗算される。

【００１５】この拡散（Ｗａｌｓｈ拡散とショートコード拡散の両方）が、広い周波数スペクトラムにデータ信号のバンド幅を広げる。このように得られたチップレート信号は、元のシンボルレート信号よりも$g=R_c/R_b$倍だけ広いバンド幅を占有する。例えば、この拡散シーケンスは、ＩチャネルとＱチャネル（ＱＰＳＫ拡散）に対しては、異なる周期的二進ＰＮシーケンス（ＰＮは疑似ノイズ）である。この拡散シーケンスは、「チップシーケンス」とも称し、そのため拡散後の処理レートは、「チップレート」R_cとも称する。

【００１６】未変調パイロット符号（Ｗａｌｓｈ符号は０で、常に＋１で、そのデータは常に＋１）がデータ信号内に組込まれ、受信機のデータチャネルのコヒーレント復調用に位相基準として用いられる。全てのユーザに対して１個のパイロットチャネルで十分であるが、それは同期ＣＤＭＡリンクだからである。

【００１７】符号化器１０４－１０５からの出力は、それぞれＦＩＲフィルタ１０６と１０７でフィルタ処理される。ＦＩＲフィルタ１０６，１０７の出力は、その後それぞれ無線キャリア周波数信号ｃｏｓ（ω_cｔ）とｓｉｎ（ω_cｔ）を用いて変調器１０８と１０９によりアップコンバートされる。変調器１０８，１０９の出力は無線周波数信号であり、これらの信号はコンバイナー１１０内で結合されて、アンテナ１１１を介して無線で移動局ユーザに送信される。

【００１８】この無線周波数ＱＰＳＫ／ＣＤＭＡ信号は、合算された全てのチャネル（データチャネル，パイロットチャネル）を含む。レートR_bでのベースバンド内（シンボルとも称する）の１ビットは、チャネル上のレートR_cのｇ個のチップからなる。

【００１９】例えば、ＩＳ－９５の送信器においては、パラメータは次の通りである。R_b＝１９．２kbps（kilobit per second），R_c＝１．２２８８Ｍｃｐｓ（megachip per second），ｇ＝６４。

【００２０】図２は、移動局で使用されるＣＤＭＡ受信機を表すブロック図である。アンテナ２０１を介して受信した無線周波数信号は、変調器２０２，２０３によりそれぞれ無線周波数信号ｃｏｓ（ω_cｔ）とｓｉｎ（ω_cｔ）を用いてダウンコンバートされる。ダウンコンバータの機能を実行する変調器２０２，２０３の出力は、それぞれアンチアリアシングＬＰＦ（ローパスフィルタ）２０４，２０５によりフィルタ処理されて、ベースバンドＩ信号とベースバンドＱ信号を生成する。

【００２１】その後、このＩ信号とＱ信号は、デジタル信号プロセッサ（ＤＳＰ）２０９の制御下で動作するＣＤＭＡＲＡＫＥ受信機２０８により復号化され、逆拡散されて、出力データ信号２１０を生成する。ＤＳＰは、異なるマルチパス成分を追跡する別々のフィンガにより受信したデータ信号の重み付き平均を出力する。

【００２２】パイロット干渉除去（pilot interference cancellation）を行う本発明のＣＤＭＡ復調器フィンガの実施例を説明する前に、従来技術にかかるＣＤＭＡ ＲＡＫＥ受信機の動作について説明する。ＲＡＫＥ受信機は、他のユーザに起因する干渉が存在しない場合には、マルチパス環境内で信号を受信するのに最適なメカニズムである。しかし本発明のＣＤＭＡシステムは、他のユーザによる干渉が存在する場合に適用される。その理由は、所望の信号と干渉信号の間の相互相関（cross-correlations）は非常に低く、ＲＡＫＥ受信機は非常に良好な（しかし必ずしも最適ではない）の性能を与えるからである。

【００２３】ＲＡＫＥ受信機の例は、次の文献に記載さ

れている。

1) "A Communication Technique for Multipath Channels" by R. Price and P.E. Green Jr.; Proceedings IRE, Vol. 46, Pages 555-570, March,1958

2) "Introduction to Spread Spectrum Anti-multipath Technique and Their Applications to Urban Digital Radio" by G. L. Turin; Proceedings IEEE, Vol. 68, No. 3, Pages 328-353, March, 1980

3) "Digital Communications" by J. G. Proakis; McGraw-Hill, 1989

【００２４】図３は、ＣＤＭＡ受信機のブロック図である。ＲＡＫＥ受信機は、マルチパス環境において、異なるパスを介して到達した受信信号の固有の時間ダイバシティの利点を利用して、ＣＤＭＡシステムの順方向リンクと逆方向リンクの両方で用いられる。

【００２５】アナログのＩ信号（Ｉ）とＱ信号（Ｑ）は、それぞれＡ／Ｄ回路３０１と３０２でデジタル信号に変換される。制御論理回路３０３は、デジタル信号プロセッサＤＳＰのインタフェース機能と、制御機能と、共通タイミング機能とをＣＤＭＡ受信機に与える。制御論理回路３０３は、ＤＳＰ（図示せず）からのＤＳＰバスを介して受信した信号の制御下で動作する。ＲＳＳＩ（受信信号強度インディケータ）３０４が、様々な信号パスを介して受信したＩ信号とＱ信号の全受信信号パワーを計算する。

【００２６】ＲＡＫＥ受信機においては、数個（通常４個）のほぼ同一機能のフィンガユニット３０５−３０８がある。各フィンガ３０５−３０８を用いて、マルチパス環境下で異なる空間パスを介して到着した受信信号を復調する。これらのフィンガ３０５−３０８は基本的に同一であるが、但しこれらは時間遅延，減衰特性，位相特性が異なる。フィンガユニット３０８はさらに付属の小型論理回路を有し、それを高速パイロットサーチャとして用いることができる（図１に示したＷａｌｓｈ信号パイロットを検出するコヒーレント受信機内で使用される）。

【００２７】パイロットサーチャフィンガ３０８は、来入信号をパイロットＰＮシーケンスでもって連続的に相関をとることにより来入信号を検査する。パイロットサーチャフィンガ３０８は、異なる基地局とマルチパス成分とを検出し、それぞれのＰＮオフセットを復調用フィンガ３０５−３０７に配信する。

【００２８】各復調器のフィンガは、来入したマルチパスで歪んだ信号のあるパスのコヒーレント復調を実行する。

【００２９】図４は、コヒーレントＣＤＭＡの従来技術にかかるレイクフィンガのアーキテクチャである。コヒーレントＣＤＭＡ受信機内のＩＳ−９５ＲＡＫＥフィンガは、３個の複合相関器を有し、それぞれ４０２はパイロットオンタイム検出用で、４０３はパイロット早期／遅延の検出用で、４０４はデータオンタイムの検出用であり、これらが一体となってタイミング信号を再構成する。この構成によりデータ複合相関器４０４による単一Ｗａｌｓｈチャネル上のデータの復号化と、逆拡散化が可能となる。相関器４０２−４０４のデータ出力は、その後ＤＳＰバスを介してＤＳＰ４２０に出力される。

【００３０】Ｉ／ＱＰＮ生成器４０５が、入力符号を相関器４０２−４０４に与える。Ｗａｌｓｈ関数生成器４０６は、Ｗａｌｓｈ符号をデータ複合相関器４０４に与える。制御回路４０７とスリュウ制御論理４０８がＲＡＫＥフィンガの動作用に制御信号を与え、かつＤＳＰバスへのインタフェースを与える。

【００３１】次に従来の変数名の定義を示す。

T_c　秒あたりのチップ持続時間

$R_c = 1/T_c$　チップレート、ＩＳ−９５では１．２２８８Ｍｃｐｓ

$R_b = R_c/N_c$　ビットレート（即ち、シンボルレート）、ＩＳ−９５では１９．２ｋｂｐｓ

N_c　シンボル（ビット）あたりのチップ数、ＩＳ−９５では６４

【００３２】Ａ　パイロットゲイン（単一のユーザ振幅と比較した）

ρ　各ＩチャネルとＱチャネルに対し、１チップ間隔の間得られたサンプル数（オーバーサンプリング係数）

$\Delta T_I = (\Delta_I + \delta_I \cdot 1/\rho)\, T_c$　メインパス成分に対するＩ番目のマルチパス成分の遅延時間、ここでΔ_1は整数部分（チップの遅延）で、δ_1がチップの分数部分（チップの１／ρ部分の遅延）、Δとδは整数値でδ＝０…ρ−

【００３３】$\tau_1 = \rho\Delta_1 + \delta_1$　サブチップ内のパス０に対する遅延；１チップはρ個のサブチップから構成される（$\tau_0 = 0$と仮定）

Ｌ　マルチパス成分の数；インデックスはｌ＝０…Ｌ−

$r_{(l)}^{(n)}$　他のマルチパス成分からのノイズを含むマルチパス成分ｌに対するｎ番目のシンボルの受信信号ベクトル（各ベクトル要素は、複素数）

【００３４】$p_{(l)}^{(n)}$　マルチパス成分ｌに対するｎ番目のシンボルのｐＮショートコード（ショートコード，パイロット符号とも称する）

$s_{k(l)}^{(n)}$　マルチパス成分ｌ（ユーザｋ）用のｎ番目のシンボルのシンボルシグネチャーコード（Walsh-code）でベクトル要素は実数

【数１】

$$\hat{c}_{(l)}^{(n)}$$

上記の符号は、マルチパス成分ｌに対するｎ番目のシンボルから得られた復号チャネル予測値（これはベクトルではない）

【００３５】

【数２】

$\hat{c}$
上記の符号は、マルチパス成分１の利用可能なチャネル予測値の組
【数３】

$$f\left(\hat{c}_{(l)}, n\right)$$

上記の符号は、より信頼できるチャネル予測値（平均化，ＦＩＲ－ＬＰフィルタ処理）を得るためのチャネル予測値上で実行される関数；計算に使用される最新の予測値はシンボルｎの予測値である
【００３６】$y_{(1)}^{(n)}$　マルチパス成分１に対するｎ番目のシンボルの復調器出力
$r^{I,Q}[i]=r[i]$　サブチップレートρR_cでの全合成のρ倍でオーバーサンプルされた復号信号
【数４】

$$\tilde{r}^{(n)}$$

上記は、受信信号ベクトルρR_c
【００３７】従来の受信機
図５にパイロットオンタイム複合相関器（図４の４０２）と、データ１オンタイム複合相関器（図４のと４０４）の基本的な復調器の構造をマルチパス成分が０の複合信号処理ブロックとして示す。要素５０１－５０４は、パイロットオンタイム相関器４０２の機能を与え、一方５０１－５０３，５０７，５０８は、データ１オンタイム相関器４０４の機能を与える。図５に用いられた複素数のグラフ表示は、同図に示すように入力信号は$r[i]=r^I[i]+r^Q[i]$である（即ち、図４のＩ信号とＱ信号である）。
【００３８】来入信号ｒ［ｉ］は、チップ－シンボルあたりのρ個のサンプルである、オーバーサンプルされた複合ＱＰＳＫＤＳ／ＣＤＭＡベースバンド信号（ダウンコンバート後の）である。オンタイムセレクタ５０１は、後続の処理のためにチップあたりρ個のサンプルのうちの１つをピックアップする。信号$r_{(0)}[i]$を乗算器５０３内で複合共役回路５０２から受信した適宜に整合化したショートコードのＰＮシーケンス$p_{(1)}^{(n)}$と乗算することによりパイロット信号の逆拡散が実行される。
【００３９】乗算器５０３からの信号から上側のアキュムレータ通路（１シンボルに亘る蓄積（accumulation））からチャネル予測値（数１式）が得られる。チャネル予測相関器と称するこの上側通路は、アキュムレータ５０４と選択的にチャネル予測アルゴリズムブロック（Channel Estimation Algorithm Block（ＣＡＬ））５０５と、複合共役回路５０６とを有する。
【００４０】特定のマルチパスに対するチャネル係数は、シンボル毎に大幅に変化するわけではないので、現在のシンボルに対するチャネル係数予測は、ＣＡＬ５０５により改善され、そしてこのＣＡＬ５０５は、アキュムレータ５０４からの現在の出力と全ての得られたチャネル予測の重み付き平均を生成する。シグネチャーコード$s_{k(1)}^{(n)}$の（ユーザｋのＷａｌｓｈ符号）を除去すると、下側通路、即ちデータ相関器５０７－５０８は二進情報を再構成し、この二進情報を乗算器５０９内で上側通路からの複素共役チャネル予測値（チャネル重み付け値）と乗算することにより、信号スペース（位相／減衰相関）に整合する。
【００４１】ブロック５１０は、乗算器５０９からの複合信号（チャネル予測出力とデータ相関器出力の積）の実部をとり、それを図６のデジタル信号プロセッサ（ＤＳＰ）６２０として示す受信機の復号化部分（ビタービ復号化器，スライサーあるいはマルチパス結合器）に入力する。
【００４２】ＣＡＬブロック５０５に関しては、チャネル予測値（数１式）は、シンボルレートでもって得られる。（このチャネル予測値は、パイロットチャネル振幅を含む。理由は、このチャネル予測値は、パイロットチャネルと相関をとることにより得られたためである。）より信頼性のあるチャネル予測値を得るためには、最後のＮα個のチャネル予測値（ｎはチャネル予測計算アルゴリズムに含まれる最新のチャネル予測値のインデックスとする）のある種の重み付けの和である下記式をとること、例えばローパスＦＩＲフィルタ処理をすることは一般的である。
【数５】

$$f\left(\hat{c}_{(l)}, n\right)$$

【００４３】ＣＡＬアルゴリズムの利点は、フェージングとＶＣＸＯオフセットのようなチャネル特性により、制限される。その理由は、チャネルパラメータは、平均化（即ち、線形挿間も可能である）の間ほとんど一定に維持しなければならないからである。ＣＡＬの複雑さを増加させることを考えると、大部分の時間１シンボルに亘るチャネル予測値で十分であることが分かる。しかし、後述するように本発明のパイロット除去系はある特定のチャネル予測アルゴリズムに限定されるものではない。
【００４４】以下の説明においては、どのチャネル予測値（最新の？）がパイロット再構成に使用されるＣＡＬに含まれるか、および復調のどの部分が除去の利点を利用しているかを明確にすることが重要である。復調は常に最新のチャネル予測値の知識を有しているが、パイロット再構成は、必ずしも有しているわけではない。
【００４５】アキュムレータブロックに関しては、その出力点で記憶容量を有している、あるいは要素をホールドできると仮定している。アキュムレータは、各シンボルクロックサイクル毎に最新の蓄積値を新たな蓄積値で更新するまでその値を保持している。
【００４６】図４の従来の受信機に基づいて、前置および後置の復調除去構造の両方に本発明のパイロット除去

系を適用した数個の実施例を示す。本発明の構造は、ブロック５０３－５１０を含む復調ユニット５２０を用いている。

【００４７】図面を単純化するために、２個のパス信号に着目し、このため２個の復調フィンガ、即ちフィンガ０とフィンガ１のみを示す。これ以上の数のパス／フィンガへの拡張は当業者には容易であろう。

【００４８】前復調除去系

前復調除去系においては、パイロット干渉除去（減算）がチップ－サンプル上で実行される。

【００４９】図６に本発明のパイロット干渉除去（Pilot Interference Cancellation（ＰＩＣ））系の外観を示す。図６の本発明の受信機は、２本のパス６１１と６１２のみを経由した信号を受信し、そのため受信機の通常は３本以上のフィンガのうち２本のみを使用する（図３参照のこと）。フィンガ６０３，６０４は、それぞれ異なるパス信号６０１と６０２を前述した方法で復調するよう動作する。

【００５０】この実施例の受信機は、複合信号$r^{(n)}$を受信するが、それがパス０とパス１からの両方の信号を表すことは認識していない。これらパス０とパス１の信号は、減衰α，位相φ，パス遅延τの点で異なっている。パイロット信号は、受信パス信号のパワーの約２０％を示すので、パス１のパイロット信号をパス０の受信信号から取り除くとができ、そしてその逆もまた可能でるならば、その結果受信機はより正確な復調を実行できることを発明者は認識した。

【００５１】このことを考慮すると、フィンガ６０３，６０４を変更してパイロット再構成回路６０６，６０７をそれぞれそれらのフィンガ６０３，６０４に追加して、パス０からのパイロット信号６１１と、パス１からのパイロット信号６１２を再構成する。パス０の受信信号$r^{(n)}$である$r^{(0)}$は、オンタイムセレクタ回路（on-time selector circuit（ＯＴＳ））によりまず処理されて、その後変更したフィンガ６０３により処理される。パス１の受信信号$r^{(n)}$である$r^{(1)}$は、同じくオンタイムセレクタ回路（ＯＴＳ）６０２によりまず処理されて、その後修正されたフィンガ６０４により処理される。

【００５２】パイロット再構成回路６０６，６０７は、予測された減衰α，位相φ，パス遅延τでもってパイロット信号を再構成する。同図に示すようにパス０からの再構成パイロット信号は、加算回路６０９内でパス１の信号から減算される（即ち除去される）。本発明の前除去系においては、パイロット干渉除去（減算）は、復調が行われる前にチップ－サンプル上で実行される。

【００５３】同様にパス１からの再構成パイロット信号は、加算回路６０８内でパス０の信号から減算される。それぞれパス１とパス０のパイロット信号を減算して得られたパス０とパス１の信号は、その後さらにそれぞれフィンガ０とフィンガ１内で正確に復調される。前述したのと同様に、フィンガ０と１からの出力信号は、例えばＤＳＰ６０５内で実行されるようなビット決定、あるいはビタービ復号化装置内で処理される。

【００５４】Ａ．－バッファを有するパイロット除去（ディテクタＡ）

図７はパイロット除去用に最新のチャネル予測を獲得するために、シンボルバッファを用いた２フィンガの前除去構成の詳細図である。本発明によれば、パイロットを再構成し、それを復調の前に除去する現行のシンボルのチャネル予測値を使用するために、データは蓄積しなければならない。その後この処理は次の３段階で行われる。

【００５５】１．各フィンガＩに対して、受信信号からＩ番目のマルチパス成分のチャネル予測値を得て、この予測値を用いてパイロットを再構成する。

２．Ｌ個の蓄積された受信信号の各々に対して、他の（Ｌ－１）個のマルチパス信号により引き起こされたパイロット干渉を除去するためにこの再構成されたパイロットを使用する。

３．このようにして得られた信号を復調する。

【００５６】図７は、この手順に従って作用する構成を示す。以下の説明においては、ダッシュを付けた番号は、ダッシュの付いていない番号のブロックと同一の働きをする。同図に示すようにブロック５０１′－５０５′，５０９′，５２０′は、ブロック５０１－５０５，５０９，５２０（図５に示す）と同じ働きをする。ブロック７００と７００′は、正規化された（７０５と７０５′により）チャネル予測値（５０４－５０５，５０４′－５０５′を用いて得られた）を用いる別個のパイロットディテクタである。

【００５７】パイロットディテクタ７００と７００′のパルス整形は、それぞれＲＬＰブロック７０１と７０１′（再構成ローパスフィルタ）を用いて考慮に入れられる。ＲＬＰ７０１と７０１′は、遅延が複数のチップ持続期間に存在しない場合には必要である。パルス整形を考慮しないとビットエラーレート（Bit Error Rate（ＢＥＲ）は増加する。ＲＬＰの実現方法を以下に説明する。

【００５８】パイロットディテクタ７０１と７０１′がパイロット信号を再構成している間、シンボルバッファ７０３－７０４と７０３′－７０４′により、シンボルデータを蓄積しておくことができる。

【００５９】一方のブランチにおいてＲＬＰにより導入された遅延を除去するために、小さなＲＬＰ遅延バッファz^{-D}が他方のブランチに付加されるために必要である（Ｄはチップ－サンプルのＲＬＰ遅延で、Ｄ＝Ｎ／２で、ＮはＲＬＰフィルタのタップ数）。かくして遅延量z^{-D}７０２と７０２′は、それぞれＲＬＰ７０１と７０１′を補償する。

【００６０】選択事項として、あるチップの範囲における遅延オフセット（シンボルあたりのチップの全数に比較して小さな）に対して、遅延７０６－７０８と７０６′－７０８′として示される下記の整合バッファの影響を考慮してもよい。

【数６】

$$z^{-\tilde{\Delta}_l}, \tilde{\Delta}_l = \left(\max_{l=0,L-1} \Delta_l\right) - \Delta_l$$

【００６１】このような影響は、無視できると我々は考えた。理由は実際に実現する際に全てのフィンガの出力の組み合わせはシンボルレートで実行されるからである。それ故に整合バッファ７０６－７０８，７０６′－７０８′は必要ではない。このことを考慮すると、チップ－サンプルレベルで整合バッファに必要とされるハードウェアは必要ではない。そのため以下に示す実施例では、この整合バッファは取り除いてある。

【００６２】パス０のパイロット信号がパイロットディテクタ７００内で再構成された後、このパイロット信号は加算器７１１″に加えられ、復調器５２０′による信号の復調の前にパス１の信号から減算される。パイロットディテクタ７００のＲＬＰブロック７０１からの非パルス整形（遅延しただけであり、ＲＬＰの説明を参照のこと）出力は、遅延され、共役化され、乗算器５０３への入力として用いられる。

【００６３】同様にパス１のパイロット信号は、パイロットディテクタ７０１′内で再構成され、これは加算器７１１に加えられ、復調器５２０による信号の復調の前にパス０の信号から減算される。パイロットディテクタ７００′のＲＬＰブロック７０１′からの非パルス整形出力は、遅延され、共役化され、そして乗算器５０３への入力として用いられる。

【００６４】本発明の他の実施例によれば、ディテクタＡを変更すると、パイロット再構成に使用されるチャネル予測値（７００と７００′から得られる）は、復調プロセス（復調器５２０と５２０′による）にも使用できるようになる。このような実施例においては、本発明はフィンガあたり２個の位相予測蓄積とＣＡＬブロックを必要とはしない。しかし、このことを実行することにより、データ相関器のみがパイロット除去から利点を受け、そのためこの構成はＢＥＲ性能が落ちる。この同一構成の若干の変更は、次に述べるディテクタＣについても適用可能である。

【００６５】図８，９には図７のディテクタＡのタイミングチャートを示す。図８のタイミングチャートは、パイロット再構成用のチャネル予測値が、シンボルタイミングに対していかに得られるかを示したものである。シンボルバッファ（例、７０３）は、シンボルｎのパイロット除去がシンボルｎのデータから得られたチャネル予測値（最後に得られたチャネル予測値）に対し実行することを補償している。整合バッファ（例、７０６）により、パイロット除去用の新たなチャネル予測値は復調プロセスの開始点で得られる。

【００６６】図９には、パイロット再構成用のチャネル予測値が整合バッファを利用しないタイミングチャートを示す。整合バッファを有しない性能の劣化は、遅延オフセットτ_1があるチップのディメンジョン内にある場合には（τ_0＝０と仮定して）無視できる。

【００６７】ディテクタＡの動作

レイレイフェージング環境においては、ディテクタＡの利点は、ディテクタがパイロット再構成用に得られる最新のチャネル予測値を有しており、これがチャネル特性がシンボル毎に大幅に代わるような場合にＢＥＲ上に好ましい影響を有することである。

【００６８】しかし、ＡＷＧＮチャネルにおいては、チャネル特徴はシンボル毎に変化しないので、検出器Ｂに対する改善点は存在しない。

【００６９】パイロット信号を再構成するために、チャネル予測値は除去段を通らない信号から得られる。この点に関しては、後述する巡回構成は利点を有するが、その理由はチャネル予測値は除去段を既に通過したデータから得られ、それ故にノイズの影響が少ないためである。

【００７０】実際にはチップ－サンプルは、４ビット解像度（ＩサンプルとＱサンプル）を有する。かくしてパイロット再構成プロセス全体は、低ビット解像度（ＲＬＰ：４ビット，チップレート乗算：４ビット）でもって動作することができる。位相整形を考慮しない場合には、チップレートでの乗算さえも必要ではない。チップレートの乗算を避ける別の方法は、後述する後変調計で説明する。

【００７１】Ｂ－バッファなしの循環型構成（ディテクタＢ）

図１０には前のチャネル予測値を用いることのない循環型前除去を使用する受信機を示す。この構成は、バッファが必要ないために実現するのに最も有望な構成である。付属のハードウェアは最少で済む。復調器５２０（フィンガ０）と５２０′（フィンガ１）から前のシンボルｎ－１の復調に用いられるチャネル予測値１００１と１００１′をそれぞれフィンガ０と１の次のシンボルｎのパイロット信号の再構成と除去に再利用する。このパイロット再構成は、回路１０１０と１０１０′で行われる。除去は、減算器７１１，７１１′内で行われる。ディテクタＢは、残りのブロックは、前述したブロックと同一の番号を付してある。

【００７２】図１１には、パイロット再構成用にチャネル予測タイミングを示すディテクタＢのタイミングチャート図を示す。シンボルｎのパイロット除去は、シンボルｎ－１のデータから得られたチャネル予測値と共に一部が動作する（シンボルｎ，ｎ－２の予測値と共に）。

【００７３】ディテクタＢの動作

この構成の不利な点は、最新のチャネル予測値が除去に利用できないことであり、その結果高速のフェージング環境において、性能が劣化することになる。

【００７４】しかし、循環ループには有効な副次的効果がある、この構成ではパイロット信号を再構成するのに用いられるチャネル予測値は、前述したパイロット除去段を通過したデータから得られる。

【００７５】Ｃーーバッファを有する循環型構成（ディテクタＣ）

図１２は、上記で説明した２つの概念を組み合わせた構成を示す。まず第１に、パイロット再構成用にＣＡＬ内で得られる（７００と７００′による）最新のチャネル予測値を有するバッファ（即ち、７０３と７０３′）を用い、そして第２に、循環ループ（即ち、１０１０と１０１０′）を用いて、その結果パイロット再構成のチャネル予測さえも前のチャネル除去から利点を得ることができる。

【００７６】同図に示すように巡回ループブロック１０１０′により、パス１からの検出パイロット信号は、（加算器１２１０を用いて）パス０のパイロット検知器７００へ入力される信号から減算される。同様に巡回ループブロック１０１０により、パス０からの検出パイロット信号は、（加算器１２１０′を用いて）パス１のパイロット検知器７００′へ入力される信号から減算される。再びディテクタＣの残りのブロックは、前述した図と同一の方法で番号がふされている。

【００７７】パイロット再構成用のチャネル予測値を使用するディテクタＣ用のタイミングチャート（図示せず）は、ディテクタＡのそれと類似する。但し、パイロット再構成用のよりよいチャネル予測値を提供するさらに別の除去手段を有する点が異なる。

【００７８】ディテクタＣの動作

ディテクタＣは、パイロット再構成用に得られる最新のチャネル予測値と前の除去から利点を有するデータから得られるパイロット再構成用のチャネル予測値の両方を組み合わせる。しかし、ディテクタＢの性能向上は、ハードウェアの複雑さ（シンボルバッファとＲＬＰを用いた第２のパイロット再構成処理とチップレートでの乗算が必要である）を正当化できるほど優れたものではない。

【００７９】後蓄積除去系

この後除去系においては、パイロット除去（減算）は、シンボルレートR_bで実行される。後除去を行う目的は、チップレートR_cでの乗算を回避するためである。

【００８０】Ａー後復調除去（ディテクタＤ）

前除去系においては、例えば図１３においては、再構成パイロット信号$C_{(1)}$［ｉ］は、復調の前にチップレートで受信信号ｒ［ｉ］から加算器１３００内から除去される。図１３のＡにおいては、除去段（加算器１３０１と１３０２）は、蓄積器（１３０３と１３０４）の後ろに配置され、シンボルレートサンプルに対し除去を実行する。

【００８１】図１４のディテクタＤは、オンタイムセレクタ５０１と５０１′，ＣＡＬ回路１４１０と１４１０′，復調器５２０と５２０′とパイロットディテクタ１４１０と１４１０′，加算回路１４２０とを有する。オンタイムセレクタ５０１と５０１′と復調器５２０，５２０′の動作は前述した通りである。

【００８２】ディテクタＤは、個々に蓄積され（パイロットディテクタ１４１０と１４１０′内で）、シンボルレートR_bでのチャネル予測値（復調器５２０と５２０′とＣＡＬ回路１４０１と１４０１′からの）で乗算された（乗算器１４０２，１４０３と１４０２′，１４０３′）再構成パイロット信号を用いる。その後１４０２と１４０２′から得られたパイロット信号は、加算器１４２１と１４２１′内で加算されて、それぞれ５２０と５２０′のトラフィック（データ）信号となる。その後乗算器１４０３と１４０３′から得られたパイロット信号は加算器１４２２と１４２２′内で加算されてそれぞれ復調器５２０と５２０′のチャネル予測値になる。

【００８３】後除去の有効な副次的効果は、パイロット再構成用にＣＡＬ内に最新のチャネル予測値を含むためにシンボルバッファ（例、図７の７０３）はもはや必要なく（そして整合バッファ、例えば７０６さえも必要ではない）、その理由はチャネル予測乗算は、現行シンボルの終わりまで遅延するからである。ディテクタＤの構成は、図７に示す整合バッファを示すディテクタＡと等価であるが、シンボルバッファも整合バッファも必要とはしていない。

【００８４】図１５にはディテクタＤのタイミングチャートが示されている。同図に示すようにシンボルレートR_bでの蓄積器の結果は、全てのパスが時間的に整合するまで保持され（遅延オフセットは、シンボル期間よりも通常短い）、そしてその後除去が最新のチャネル予測値で実行される。シンボルレートでの遅延整合制御（蓄積器の出力が保持される）は、同図には明示していない。ディテクタＤの性能はディテクタＡのそれと類似である。

【００８５】Ｂー多段後除去（ディテクタＥ）

ディテクタＤの後除去系は、前のパイロット除去の利点を利用するチャネル予測値を用いてパイロット信号の除去を行わなかった。ディテクタＥの構成を図１６に示す。

【００８６】ディテクタＥは、オンタイムセレクタ５０１と５０１′，ＣＡＬ回路１４０１と１４０１′，復調器５２０と５２０′，パイロットディテクタ１４１０と１４１０′，加算回路１４２０，乗算器１４０２，１４０３と１４０２′，１４０３′とを有し、これらの動作は前述した通りである。さらにまたディテクタＥは、ＣＡＬ回路１６０１と１６０１′と乗算器１６０２と１６

０２′とを有し、１４１０と１４１０′のチャネル予測値のみが更新される（refined） ような第１除去段を提供する。その後、より良好なチャネル予測値を用いて実際の除去が加算回路１４２０内で行われる。

【００８７】ディテクタＥのタイミンググラフは、図１５のディテクタＤのそれと同一である。

【００８８】ディテクタＥの動作

ディテクタＥの動作は、ディテクタＣに類似する（若干良好であるが）、その理由はパイロット再構成用に使用される、あるいは除去前の全てのチャネル予測値は最新のものだからである。より良好なチャネル予測値を得るためには何段の除去段でも用いることができる。しかし、最も可能性のあるものとしては１段ではハンドセット受信機（図２）内で実現するのに値しない、その理由は、若干良好な程度のチャネル予測値の利点は、それほど重要ではないからである。

【００８９】Ｅ－３フィンガ構成の例

図１７には、前のチャネル予測値（即ち、図１０のディテクタＢ）と働くバッファを有さない循環前除去を用いた３フィンガ（３パス）の受信機の構成を示す。ＩＳ－９５のハンドセット受信機では３フィンガの設計のものが提案されている。

【００９０】ＲＬＰ（例、６０１）は２個の出力を有するが、その理由はこの２個の他のフィンガ１，２は、そのオンタイムサンプルに対し異なるタイミングを有することがあるからである。かくして、例えばパイロット０のパルス整形の再構成は、フィンガ１と２に対し、２個の異なる部分遅延オフセットδ_1，δ_2を必要とする。フィンガ０（パス０）においては、フィンガ１，２（パス１，２）の両方からのパイロットは、フィンガ０に入力される信号から除去される。同様にフィンガ１，２もその入力信号から除去された他のチャネルのそれぞれのパイロット信号を有する。この受信機の残りの部分は図１０のＢで説明したディテクタと同一の動作をする。

【００９１】再構成ローパスフィルタ（ＲＬＰ）

大部分の時間T_c（$\delta_1 \neq 0$）の小数点以下の遅延となるマルチパス成分が存在する。その後、パルス整形が考慮に入れられる。

【００９２】Ａ－パルス形成の再構成の必要性

図１８－２０は、マルチパス成分のタイミングオフセットがT_cのマルチプル内に存在しない場合に、パルス整形再構成ローパスフィルタ（ＲＬＰ）の必要性を示している。マルチパス成分０のサンプル化パイロット信号（例、Ｉチャネル）の一部を示す。この実施例においては、サンプリング時間は、$T_s = i \cdot T_c / \rho$と仮定し、ここでｉは整数で、Ｉ／Ｑ位相シフトは存在せず、信号はオンタイムサンプルで１に正規化されていると仮定している。

【００９３】

【数７】

$$pilot_0^I[i] = P_0^I[i] * h[i] = \sum_{j=-\infty}^{\infty} P_0^I[j] \cdot h[i-j] \qquad (*1)$$

正規化されたナイキストパルスシェープでは、

【数８】

$$h[i] = \mathrm{sinc}\left(\pi \tfrac{i}{\rho}\right) \cdot \frac{\cos\left(\pi \alpha_{off} \tfrac{i}{\rho}\right)}{1-\left(2\alpha_{off} \tfrac{i}{\rho}\right)} = \begin{cases} 1, & \tfrac{i}{\rho} = 0 \\ 0, & \tfrac{i}{\rho} \neq 0 \quad (\mathrm{ISI-free}) \end{cases} \qquad (*2)$$

ロールオフ係数α_{off}（ＩＳ－９５では$\alpha_{off} \fallingdotseq 0$）で、ｉはサブチップサンプルインデックスである。

【００９４】図１８には、サブチップインデックスを有する信号Ｐ［ｉ］のショートコードシーケンスを示す。図１９には、最大４個のサイドローブのナイキストレイズドコサインフィルタ（Nyquist-raised cosine filter）の正規化された時間領域インパルス応答を示す。

【００９５】図２０には、マルチパス成分０のパルス整形パイロットの例を示す。上記のパイロット信号は、マルチパス成分０に属する。その後チップレートでのオンタイムサンプルは、（理想的には）＋１または－１（正規化された）のいずれかである、その理由は、ナイキストパルス整形フィルタは、送信機内で用いられるからである（オンタイムサンプルには近傍インパルスのＩＳＩは存在しない）。実際には、送信機内に二乗ルートのナイキストレイズフィルタ（square-root Nyquist raised filter） が存在する受信機内のパルス整形マッチドフィルタ（また、二乗ルートのナイキストレイズドフィルタ）と共に受信機のベースバンド内でナイキストレイズドコサインパルス整形が得られる。

【００９６】このパイロット信号を他のマルチパス成分（フィンガ）、例えば成分１から除去するために、それぞれのマルチパス成分１のオンタイムサンプルにおいて、パイロット信号０のパルス整形を考慮する必要がある。言い換えると、マルチパス成分０のパイロット信号を成分１から除去するためには、信号１（ＲＬＰ係数α_{J, δ_1}） のオンタイムサンプルで、パイロット０のパルス整形を再構成する必要がある。T_cのマルチプル内に遅延を仮定していないので、成分１のオンタイムサンプルはどこか（in between）にあり（このことは$\delta_1 \neq 0$を意味する）、それ故に（＊１） によればパイロット信号０のパルス整形は重要である。

【００９７】再構成ローパスフィルタのＦＩＲでの実現方法は、極めて単純である。Ｎタップ（Ｎは偶数）の有限数でもって、式（＊１） の離散畳み込み加算を近似する。

【００９８】図２１には、再構成ローパスフィルタ（ＲＬＰ）のＦＩＲ実現を示す。ＦＩＲ係数は、$\alpha_{J, \delta} =$ ｈ

［（ｊ－Ｎ／２）・ρ＋δ］で、それ故に遅延τ＝ρΔ＋δの分数部分δに依存している。δ＝０の場合には、$\alpha_{N/2,0}$ ＝のみで、他の係数はゼロとなる。これはT_cの整数倍数（integer multiples）の遅延についてあてはまる。チップレートでのパルス整形出力は次式で表される。

【数９】

$$pilot_{\delta}^{I,Q}[i]=\sum_{j=0}^{N-1}a_{j,\delta}\cdot p^{I,Q}[i-j]$$

ここでｉはチップサンプルインデックスで、$p^{I,Q}$［ｉ］はショートコードシーケンスである。

【００９９】<u>ＲＬＰを実現する他の側面</u>

１．タップの非常に小さな数Ｎ（４さらにまた２でさえ）もパイロットパルス整形の十分な近似を得るのに十分であることが分かる。

２．乗算器（係数）は単純なスイッチである、その理由は来入ＰＮシーケンスは、＋１と－１からのみ成立しているからである。

３．パルス整形ｈ［ｉ］は、Ｎ／２・ρの値に対する（対称の）ルックアップテーブルとして記憶できる。かくしてＮ＝４タップ，ρ＝８と４ビットの値に対しては、テーブルのサイズは６４ビットである。

【０１００】４．１個のルックアップテーブル（そしてスイッチも加算器も不必要）として実現できる。来入ＰＮシーケンスのＮ個の二進値を通り、少数点以下の遅延（fractional delay）δ＝０‥ρ－１に従って出力を生成する。このテーブルのサイズは、2^N・ρ の値であり、対称性を利用して１／４に縮めることができる。しかし、より複雑なアクセスメカニズムが必要となる。そして再びＮ＝４タップで、ρ＝８と４ビットの値の場合には、テーブルサイズは対称性を利用しない場合には５１２ビットで、対称性を利用した場合には１２８ビットである。最も可能性のあるものとして３が実現し易い。

【０１０１】５．ＦＩＲフィルタにより導入されたチップレベルでの遅延処理は、D_{FIR} ＝Ｎ／２チップである。この遅延を補償するために、ＰＮショートコードのＲＬＰへの入力は、ＰＮシーケンスを復調することに比較してＤ個のチップだけ前もって実行される（ＰＮ周期内で）。このことは、タップ付きの遅延ラインの中央部から復調用のＰＮシーケンスを取り出すことにより容易に実行できる（図２１のポイントＸを参照のこと）。

【０１０２】６．フィンガあたり再構成する２個以上のパイロット信号が存在する（マルチパス成分１，２に対し、２個の異なる少数点以下遅延オフセットδ_1，δ_2でパイロット０を再構成する）場合には、それぞれの遅延オフセットδ_2に従ってフィンガ０で第２のＲＬＰを必要とする。ＲＬＰフィルタのタップ付き遅延ラインは、両方にとって同一であるため、新たな組の係数α_{j,δ_2}を既存のＲＬＰα_{j,δ_1}に加え、タップ付き遅延ラインを共有する必要がある。これにより複雑さが解消する。

【０１０３】<u>さらなる実現方法</u>

高速フェージングのシミュレーションにおいては、パイロット干渉除去は、パイロット信号再構成用に使用されるチャネル予測値が所定のパワーしきい値を超えたマルチパス成分から得られる場合にのみ、パイロット干渉除去を実行しなければならないことが分かった。それ以外の場合には、悪いチャネル予測値を用いることによりパイロット除去のＢＥＲ利点を不必要に損なうことになる。

【０１０４】この目的のために単純なスイッチが提案されている設計の各フィンガに付加され、その成分の受信信号パワーが小さすぎる場合（短く深いフェージングに起因して）には、それぞれのマルチパス成分用のパイロット除去を切り離す。各マルチパス成分の信号パワーは、実際の実現方法でいずれにしても計算できる。かくして、余分のハードウェアは必要とされないが、但しスイッチとしきい値検出器は必要である。

【０１０５】本発明の他の特徴によれば、スイッチは特定のパイロット信号用のパイロット干渉除去を行うか否かを制御する。このスイッチの決定は、最少平均二乗誤差基準（minimum mean-squared error criterion（ＭＭＳＥ））に従うと最適であり、線形結合器としきい値装置を用いて実現できる。単純な決定デバイスでは、キャンセルするパイロットの最適な組を決定し、理論的に本発明のパイロット干渉除去システムの性能を向上させる。

【０１０６】図２２は、決定ユニット２２０３により制御される付加切り換え機能（２２０１，２２０２）を有する図６の改善型ＰＩＣディテクタあるいは受信機を示す。下記の式をシンボル期間ｎの間、１番目のフィンガに対し除去されたパイロットの組とする。

【数１０】

【０１０７】パイロット１$\hat{G}^{(n)}$は、パス１のチャネルを予測するために必要であるため、フィンガ１から除去することはできない。そのため数１０式は、｛０，１，…，ｊ，…，（Ｌ－１）；ｊ≠１の組のサブセットである。（例えば、Ｌ＝３の場合には、除去セットの可能なグループは、下記式である。）

【数１１】

$$\hat{G}_0^{(n)}=\{1,2\},\quad \hat{G}_1^{(n)}=\{0,2\},\quad \hat{G}_2^{(n)}=\{0\}$$

【０１０８】次に述べる基準を用いて、下記のチャネル予測を用いると、

【数１２】

$$\hat{c}_0^{(n)}\qquad\text{(a)}$$

$$\hat{c}_1^{(n)}\qquad\text{(b)}$$

決定ユニット２２０３は、次のシンボル間隔で下記のパイロット除去組を決定する。

【数１３】

下記の場合には、ト$\hat{G}_0^{(n+1)}$スイ$\hat{G}_1^{(n+1)}$チはオンで、パイロット０は数１２（ａ）を用いて再構成され、次のシンボル間隔でフィンガ１への入力から除去される。

【数１４】

【０１０９】それ以外$\hat{G}_1^{(n+1)}$で下記の(0)場合には、スイッチ２２０１はオフとなり、フィンガ１に対してはパイロット０の信号除去は発生しない。

【数１５】

同様に下記の場合には、$\hat{G}_0^{(n+1)}$スイッチはオンとなり、パイロット１は数１２（ｂ）を用いて再構成され、次のシンボル期間でフィンガ１への入力から除去される。

【数１６】

それ以外に数１５式の$\hat{G}_0^{(n+1)}$場合には(1)、スイッチ２２０３はオフとなり、フィンガ０に対するパイロット１信号除去は発生しない。

【０１１０】図２３は、Ｌ＝３フィンガのＲＡＫＥ受信機の実現方法を示す。図２３は、決定ユニット２３０１とスイッチ２２０２，２２０４が付加された図１７である。パイロットを除去しようとしているフィンガのマルチパス遅延を適合するために、個別の再構成ローパスフィルタ（ＲＬＰＦ）を用いて、パイロット１の（Ｌ－１）個のバージョンを再構成しなければならない。

【０１１１】例えば、パス０に対しては、パス１と２用のパイロット信号（１７１０′と１７１０″内で）を再構成し、その後これらはパス０の信号から減算される（加算器２３０５を用いて）。同様にパイロット０と１は、パス２の信号から（加算器２３０７を用いて）減算され、パイロット０と２は、パス１の信号から（加算器２３０６を用いて）減算される。

【０１１２】スイッチ機構の変形例

パイロット干渉除去用にＭＭＳＥスイッチセットの変形例を示す。この目的はＲＡＫＥフィンガ出力の和の平均二乗エラーを最少にする下記のスイッチセットを決定することである。

【数１７】

フィンガ１の出力は、$\hat{G}_l^{(n)}$下記の$\hat{G}_l^{(n)}$スイッチセットの関数である。

【数１８】

$$y_l^{(n)}\left(\hat{G}_l^{(n)}\right)$$

【０１１３】この目的は、次の式を評価することである。

【数１９】

$$\left\{\hat{G}_0,\cdots,\hat{G}_{L-1}\right\}=\arg\min_{G_0,\cdots,G_{L-1}} E\left[\left\{\sum_{l=0}^{L-1} y_l(G_l)-\sum_{l=0}^{L-1}\bar{y}_l\right\}^2\right] \qquad (3)$$

ここで下記であり、そしてランダム符号，干渉データビット，背景熱ノイズに関して予測をとる。

【数２０】

$$\bar{y}_l \equiv E\left(y_l(G_l)\right)$$

（３）式の右側を評価すると、次のようになる。

【数２１】

$$\begin{aligned} E\left[\left\{\sum_{l=0}^{L-1} y_l(G_l)-\sum_{l=0}^{L-1}\bar{y}_l\right\}^2\right] &= E\left[\left\{\sum_{l=0}^{L-1}\left(y_l(G_l)-\bar{y}_l\right)\right\}^2\right] \\ &= \sum_{l=0}^{L-1} E\left[\left(y_l(G_l)-\bar{y}_l\right)^2\right] + \underbrace{\sum_{l=0}^{L-1}\sum_{j\ne l} 2E\left[\left(y_l(G_l)-\bar{y}_l\right)\left(y_j(G_j)-\bar{y}_j\right)\right]}_{0} \\ &= \sum_{l=0}^{L-1} Var\left(y_l(G_l)\right) \end{aligned}$$

【０１１４】和の分散（variance）は、分散の和であるので、（３）式の元の決定ルールは、次のようになる。

【数２２】

$$\hat{G}_l^{(n)} = \arg\min_{G_l^{(n)}} Var\left[y_l^{(n)}\left(G_l^{(n)}\right)\right] \quad l=0,\cdots,L-1 \qquad (4)$$

ＭＭＳＥセットは、式（５）で表される。

【数２３】

$$\hat{G}_l^{(n)} = \left\{ j:\ \left\|c_j^{(n)}\right\|^2 \ge Var\left[\hat{c}_j^{(n-1)}\left(\hat{G}_j^{(n-1)}\right)\right], j \ne l \right\} \qquad (5)$$

【０１１５】実際のチャネルパラメータ $c_j^{(n)}$ は不明であるので、決定に際しては下記の予測値を使用しなければならない。

【数２４】

$$\hat{c}_l^{(n)}\left(\hat{G}_l^{(n)}\right)$$

しかし、この置き換えは、それ自身の数１０式に依存することになる。このような状況を修復するために、シンボル間のチャネル変動は小さく、したがって式（６）の除去セットを与えるために、下記式（ｂ）の代わりに下記式（ａ）を用いる。

【数２５】

$$\hat{c}_l^{(n-1)}\left(\hat{G}_l^{(n-1)}\right) \qquad (a)$$

$$\hat{c}_l^{(n)}\left(\hat{G}_l^{(n)}\right) \qquad (b)$$

【数２６】

$$\hat{G}_l^{(n)} = \left\{ j: \left\|\hat{c}_j^{(n-1)}\left(\hat{G}_j^{(n-1)}\right)\right\|^2 \ge Var\left[\hat{c}_j^{(n-1)}\left(\hat{G}_j^{(n-1)}\right)\right], j \ne l \right\} \qquad (6)$$

【０１１６】パイロットｊの除去の決定は、対応する予測された下記のチャネルパワーと、チャネル予測値の変動にのみ依存する。

【数２７】

$$\left\|c_j^{(n-1)}\left(\hat{G}_j^{(n-1)}\right)\right\|^2$$

この決定の背景にある直感は明かである。パイロット信号ｊのパワーは、予測値の変動よりも強くなると、$c_j^{(n)}$ の基づいた再構成パイロット干渉は、信頼性が十分高く、その結果フィンガ入力からそれを除去することは、出力ＭＳＥを低下させる。

【０１１７】それ以外にパワーが弱すぎる場合には、再構成されたパイロット干渉を除去することは出力ＭＳＥを実際に増加させる。ｊ≠１という条件を除いて、この決定はフィンガ１，パイロット干渉除去用の目標フィンガには依存しない。したがって、数１０式の組は１の関数ではないが、但し１は数１０式のメンバではない、その理由はパイロットＩをフィンガ１から除去することができないからである。

【０１１８】この例外を考えると、数１７式は、下記で表すことができる。

【数２８】

$$\hat{G}_l^{(n)}$$

前に説明した実施例での除去セット数１１式は、式（５）または（６）によれば許されない。その理由はパイロット１は、フィンガ０からは除去されるが、フィンガ２からは除去されないからである。

【０１１９】除去セットの有効グループは、次式である。

【数２９】

$$\hat{G}_0^{(n)} = \{2\},\quad \hat{G}_1^{(n)} = \{0,2\},\quad \hat{G}_2^{(n)} = \{0\}$$

これらの組は、下記を用いて表すことができ、

【数３０】

$$\hat{G}^{(n)} = \{0,2\}$$

そして次式を規定することにより、

【数３１】

$$V\left(l, n, \hat{G}^{(n)}\right) = Var\left[\hat{c}_l^{(n)}\left(\hat{G}^{(n)}\right)\right]$$

次のように示される。

【数３２】

$$V(l, n-1, \hat{G}^{(n-1)}) = \frac{1}{N}\left[\sum_{j \ne l,\, j \in G_l^{(n)}} V(j, n-2, \hat{G}^{(n-2)}) + \sum_{j \ne l,\, j \notin G_l^{(n)}} \left\|c_j^{(n-1)}\right\|^2 + \frac{KA_1^2}{A_0^2}\sum_{j \ne l}\left\|c_j^{(n-1)}\right\|^2\right] + \frac{2\sigma^2}{A_0^2} \qquad (7)$$

【０１２０】ここで、Ｎは拡散係数（ＩＳ－９５ではＮ＝６４）で、Ｋは活性データ／同期チャネルの数、$2\sigma^2$ はチップあたりの熱ノイズパワー、A_0 はパイロット振幅で、A_1 はＫ個のデータ／同期チャネルの各々の振幅である。シンボル間隔（ｎ－２）からのチャネル予測値をシンボル間隔（ｎ－１）からの実際のチャネルパラメータの代わりに用いることにより式（７）は次の式となる。

【数３３】

$$V(l, n-1, \hat{G}^{(n-1)}) = \frac{1}{N}\left[\begin{array}{l}\displaystyle\sum_{j \ne l,\, j \in G_l^{(n-2)}} V(j, n-2, \hat{G}^{(n-2)}) + \\ \displaystyle\sum_{j \ne l,\, j \notin G_l^{(n-2)}} \left\|\hat{c}_j^{(n-1)}\left(\hat{G}^{(n-1)}\right)\right\|^2 + \frac{KA_1^2}{A_0^2}\sum_{j \ne l}\left\|\hat{c}_j^{(n-1)}\left(\hat{G}^{(n-1)}\right)\right\|^2\end{array}\right] + \frac{2\sigma^2}{A_0^2} \qquad (8)$$

【０１２１】各シンボル間隔において除去セット数１０式（１＝０…Ｌ－１）は、次のステップを用いて決定できる。

・式（８）を用いて１＝０…Ｌ－１に対し、下記チャネ

ル予測変動を計算する。

【数３４】

・式（６）を用いて数４１の式$\hat{G}^{(n-1)}$（l＝０…Ｌ－１）を決定する。

【０１２２】以上述べたようにパイロット干渉除去検出器用のスイッチ機構の変形例は、最少平均二乗エラーのＲＡＫＥ検出器の出力を提供できる。このスイッチに対する決定ルールは、パスｌ（ｌ＝０…Ｌ－１）のチャネル予測値のパワーが高い（即ち、その予測値の変動よりも大きい）時には、このチャネル予測値を用いて再構成された関連パイロット信号は信頼性があり、他の（Ｌ－１）個のＲＡＫＥフィンガ入力から除去しなければならない。チャネル予測値のパワーが低い場合には、再構成されたパイロット信号は信頼性がなく、このパイロットを用いた除去を行ってはならない。

【０１２３】

【発明の効果】本発明のＣＤＭＡ受信機は、Ｗａｌｓｈ符号パイロット周波数とＷａｌｓｈ符号化を用いてコヒーレント動作を与えるように順方向リンクで使用する例を用いて記載したが、コヒーレント動作を維持するような他の公知の符号系列をＣＤＭＡ送信器とＣＤＭＡ受信機（通常順方向リンク）の両方に用いることもできる。さらに本発明のコヒーレント受信機は、コヒーレント順方向リンクを例に説明したが、コヒーレント逆方向リンクにも使用することができる。

【図面の簡単な説明】

【図１】本発明の動作を説明するＣＤＭＡ通信リンクの送信器を示す図

【図２】移動局で使用されるＣＤＭＡ受信機のブロック図

【図３】ＣＤＭＡ受信機のブロック図

【図４】コヒーレントＣＤＭＡの従来技術にかかるレイクフィンガのアーキテクチャ

【図５】パイロットオンタイムとデータ１オンタイムの複合相関器（complex correlator）の基本的な復調器の構造を示す図

【図６】２フィンガのコヒーレントＣＤＭＡ受信機に適用される本発明のパイロット除去系のブロック図

【図７】本発明による前復調除去系の第１実施例（ディテクタＡ）を表す図

【図８】シンボルタイミングに関し、いかにパイロット信号再構成用のチャネル予測値が得られるかを表す図７のディテクタＡ用のタイミングチャート

【図９】シンボルタイミングに関し、いかにパイロット信号再構成用のチャネル予測値が得られるかを表す図７のディテクタＡ用のタイミングチャート

【図１０】バッファを処理しない循環前除去を用いたディテクタＢを表す図

【図１１】図１０のディテクタＢ用のタイミングチャート

【図１２】バッファ処理する循環前除去を用いたディテクタＣを表す図

【図１３】本発明による前除去系を表す図で、Ａは後除去系を表す図

【図１４】後除去を用いたディテクタＢのブロック図

【図１５】図１４のディテクタＤ用のタイミングチャート

【図１６】多段後除去を用いたディテクタＥのブロック図

【図１７】３フィンガコヒーレントＣＤＭＡ受信機に適用される本発明のパイロット信号除去系（ディテクタＢによる）のブロック図

【図１８】パルス整形（pulse-shape） 再構成ローパスフィルタ（Reconstruction Low Pass Filter（ＲＬＰ））の必要性を示す図

【図１９】パルス整形再構成ローパスフィルタ（ＲＬＰ）の必要性を示す図

【図２０】パルス整形再構成ローパスフィルタ（ＲＬＰ）の必要性を示す図

【図２１】ＲＬＰのＦＩＲ実現手段を表す図

【図２２】本発明により切り換え可能なパイロット干渉除去手段を含む２フィンガコヒーレントＣＤＭＡ受信機を表す図

【図２３】本発明により切り換え可能なパイロット干渉除去手段を含む３フィンガコヒーレントＣＤＭＡ受信機を表す図

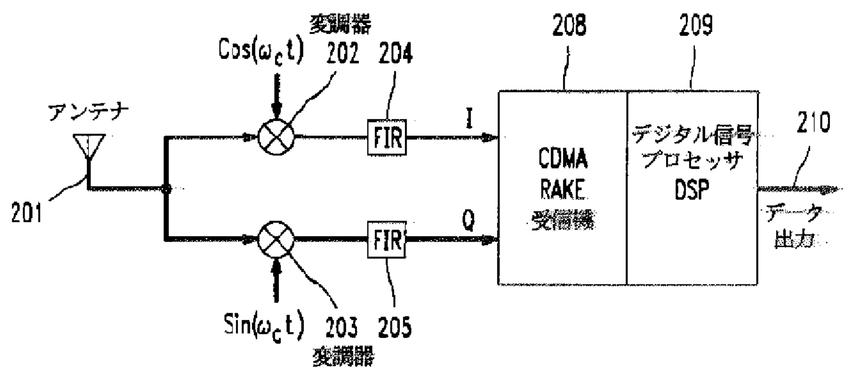
【図２４】多段後除去を用いたディテクタＥのブロック図

【符号の説明】

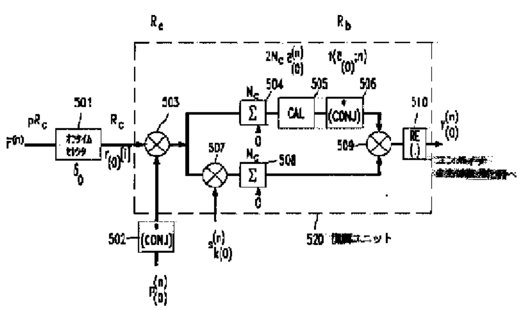
１０１　符号拡散器
１０２　加算器
１０４－１０５　符号化器
１０６，１０７　ＦＩＲフィルタ
１０８，１０９，２０２，２０３　変調器
１１０　コンバイナー
１１１，２０１　アンテナ
２０４，２０５　アンチアリアシングＬＰＦ（ローパスフィルタ）
２０８　ＣＤＭＡＲＡＫＥ受信機
２０９　デジタル信号プロセッサ（ＤＳＰ）
２１０　出力データ信号
３０１，３０２　Ａ／Ｄ回路
３０３　制御論理回路
３０４　ＲＳＳＩ（受信信号強度インディケータ）
３０５－３０８　フィンガ
４０２，４０３，４０４　データ複合相関器
４０５　Ｉ／ＱＰＮ生成器
４０６　Ｗａｌｓｈ関数生成器
４０７　制御回路

４０８　スリュウ制御論理
５０１　オンタイムセレクタ
５０２　複合共役回路
５０３，５０７，５０９　乗算器
５０４，５０５，５０６　上側通路
５２０　復調ユニット
６０１，６０２　パス信号
６０３，６０４　ＲＡＫＥフィンガ
６０５　ビターービ復号化
６０６，６０７　パイロット再構成回路
６０８，６０９　加算回路
６１１，６１２　パイロット信号
７００　パイロットディテクタ
７０１　ＲＬＰ
７０２　遅延量
７０３，７０４　シンボルバッファ
７０６，７０８　整合バッファ
７１１　加算器
１３００，１３０１，１３０２，１４２１，２３０５ 加算器
１３０３，１３０４　蓄積器
１４０１　ＣＡＬ回路
１４０２，１４０３　乗算器
１４１０　パイロットディテクタ
１４２０　加算回路
２２０１，２２０２　付加切り換え機能
２２０３　決定ユニット

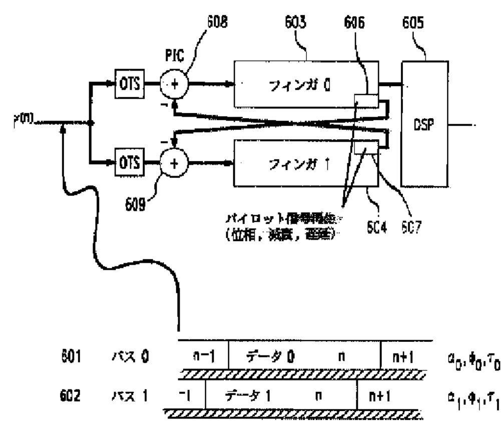
【図２】



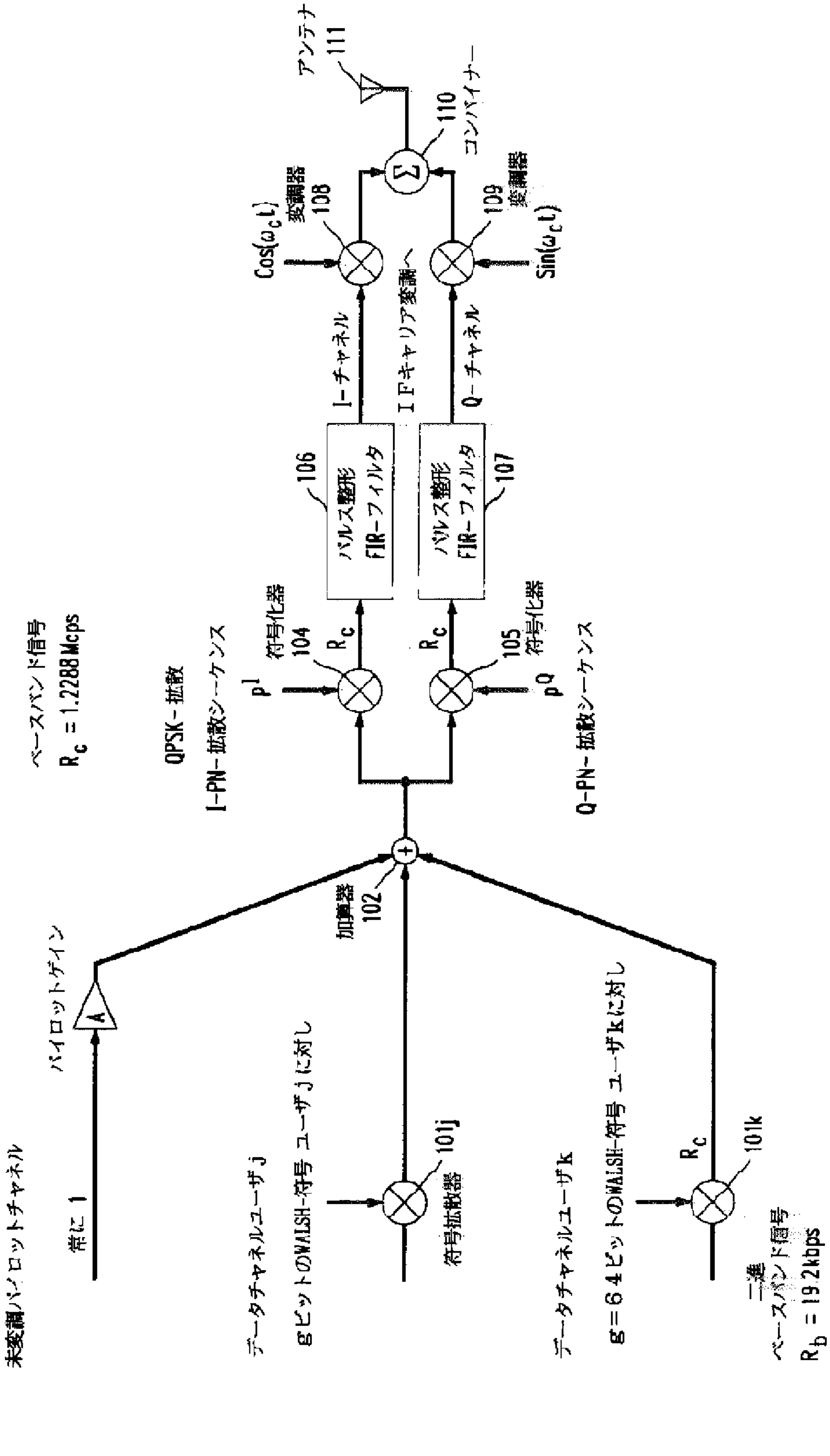
【図５】



【図６】

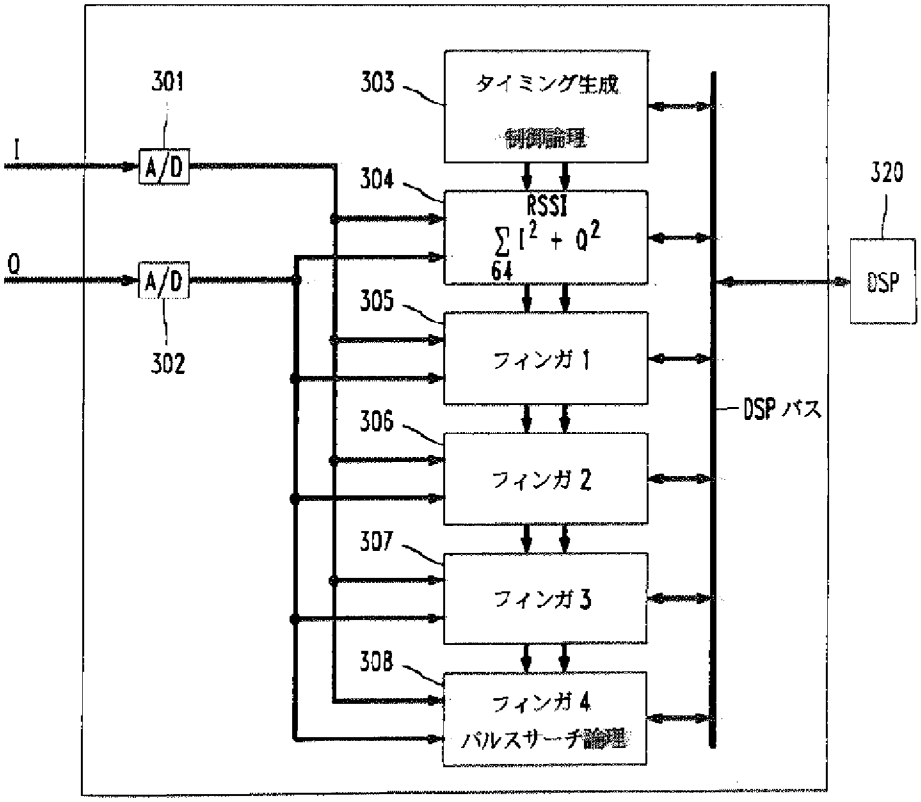


【図１】

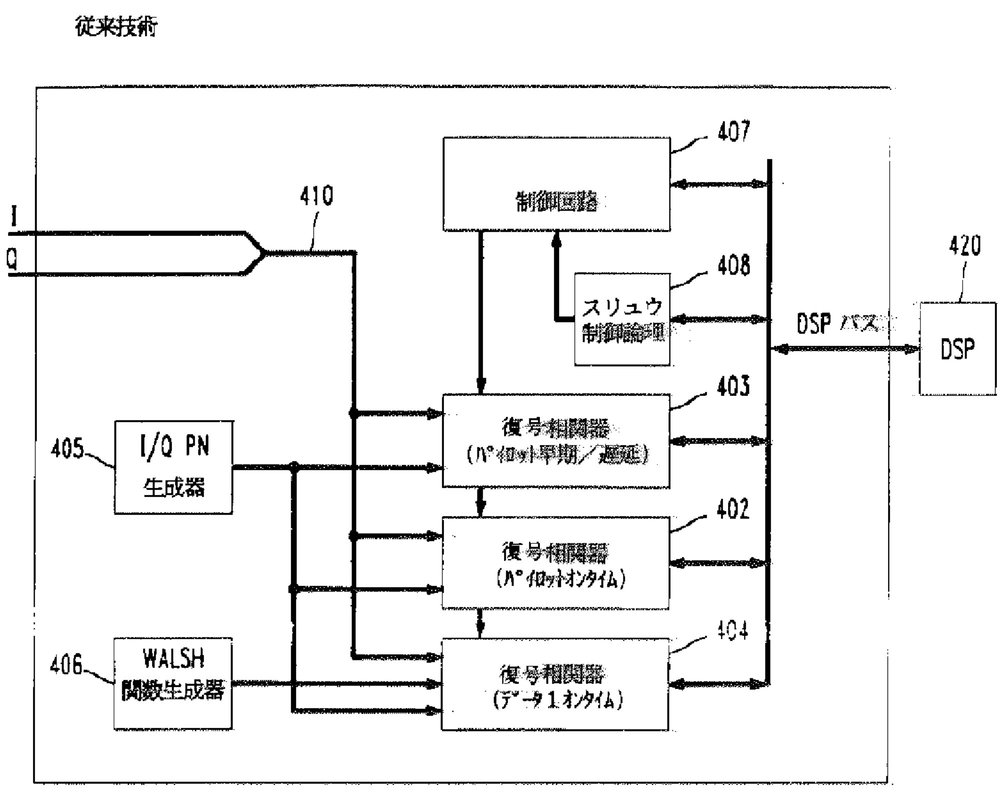


未変調パイロットチャネル
常に 1
パイロットゲイン
A
データチャネルユーザ j
gビットのWALSH-符号 ユーザ j に対し
101j
符号拡散器
データチャネルユーザ k
g＝６４ビットのWALSH-符号 ユーザ k に対し
R_c
101k
二進
ベースバンド信号
R_b = 19.2kbps
加算器
102
ベースバンド信号
R_c = 1.2288 Mcps
QPSK - 拡散
I-PN- 拡散シーケンス
Q-PN- 拡散シーケンス
p^I
p^Q
符号化器
104
105
符号化器
R_c
R_c
パルス整形
FIR-フィルタ
106
パルス整形
FIR-フィルタ
107
I-チャネル
ＩＦキャリア変調へ
Q- チャネル
Cos(ω_ct)
108
変調器
Sin(ω_ct)
109
変調器
Σ
110
コンバイナー
111
アンテナ

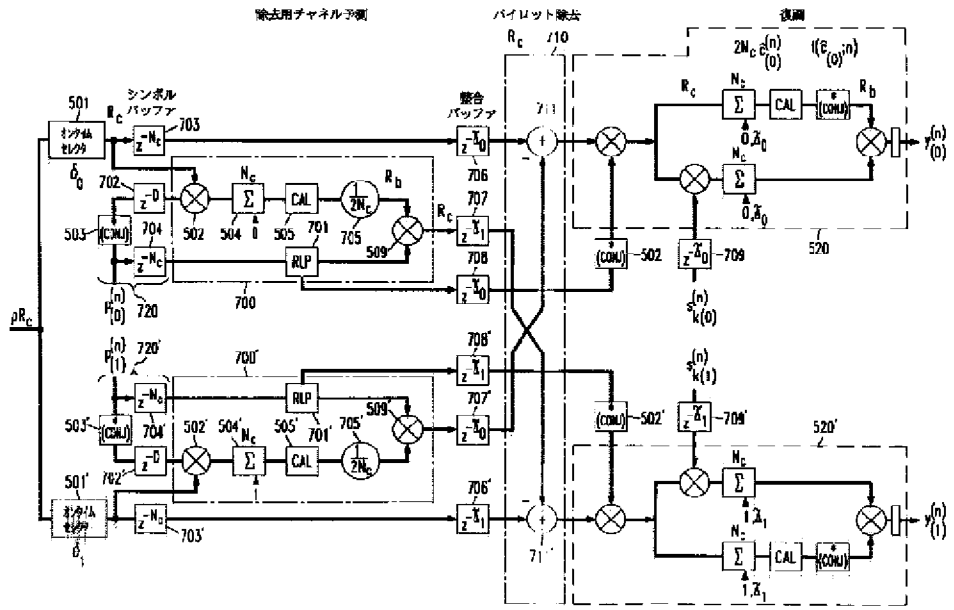
【図３】


301
303
タイミング生成
制御論理
I
A/D
304
RSSI
$\sum_{64} I^2 + Q^2$
320
Q
A/D
DSP
305
302
フィンガ 1
306
DSP バス
フィンガ 2
307
フィンガ 3
308
フィンガ 4
パルスサーチ論理

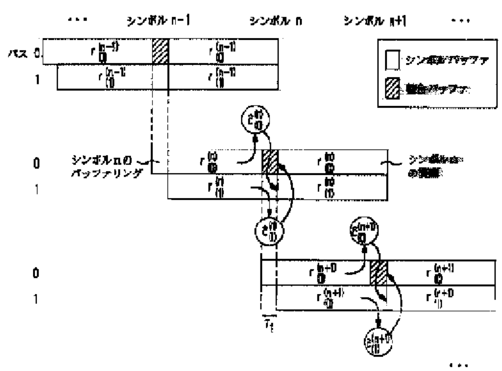
【図４】


従来技術
407
制御回路
410
I
Q
408
420
スリュウ
制御論理
DSP バス
DSP
403
405
I/Q PN
生成器
復号相関器
(パイロット早期／遅延)
402
復号相関器
(パイロットオンタイム)
404
406
WALSH
関数生成器
復号相関器
(データ1オンタイム)

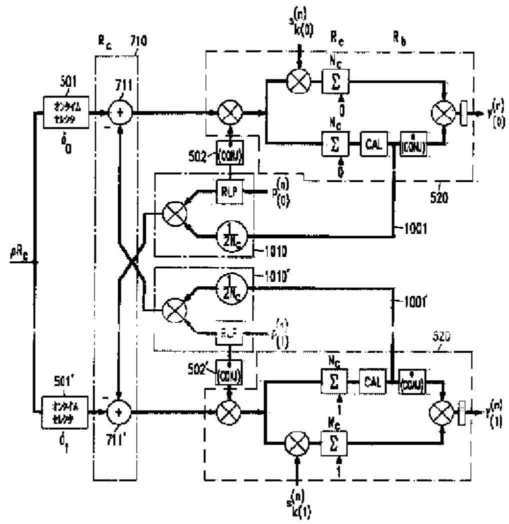
【図７】



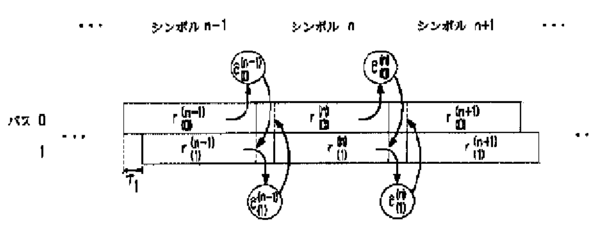
【図８】



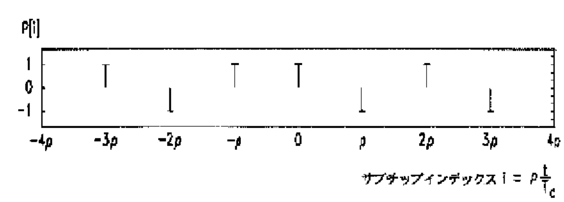
【図１０】



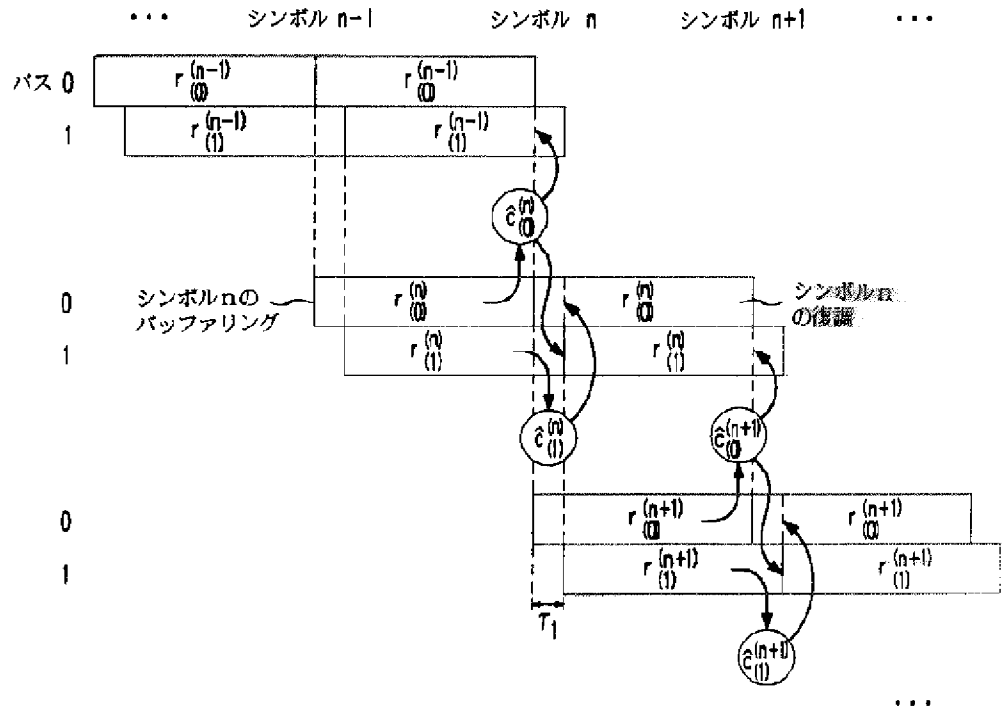
【図１１】



【図１８】

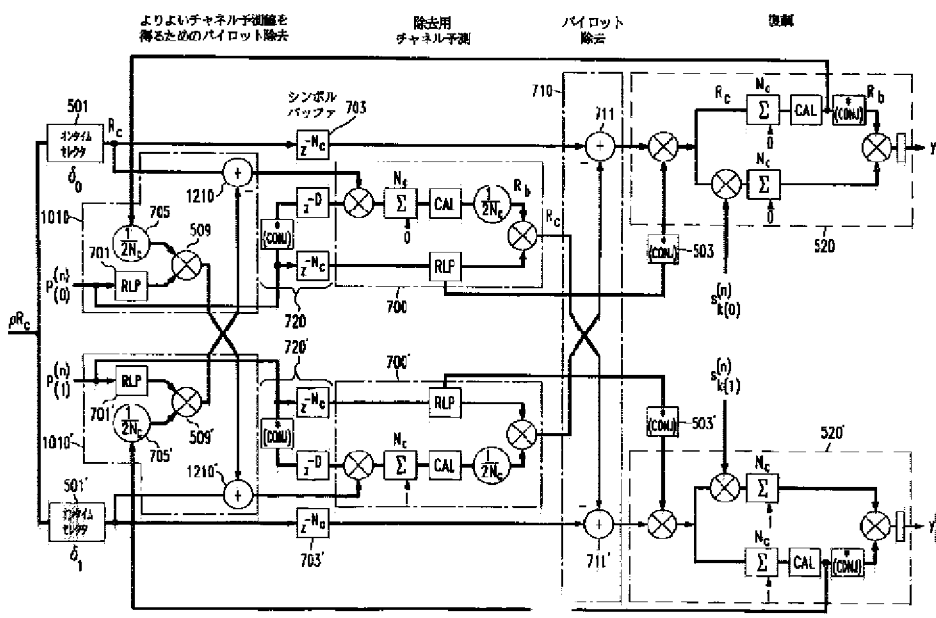


【図９】



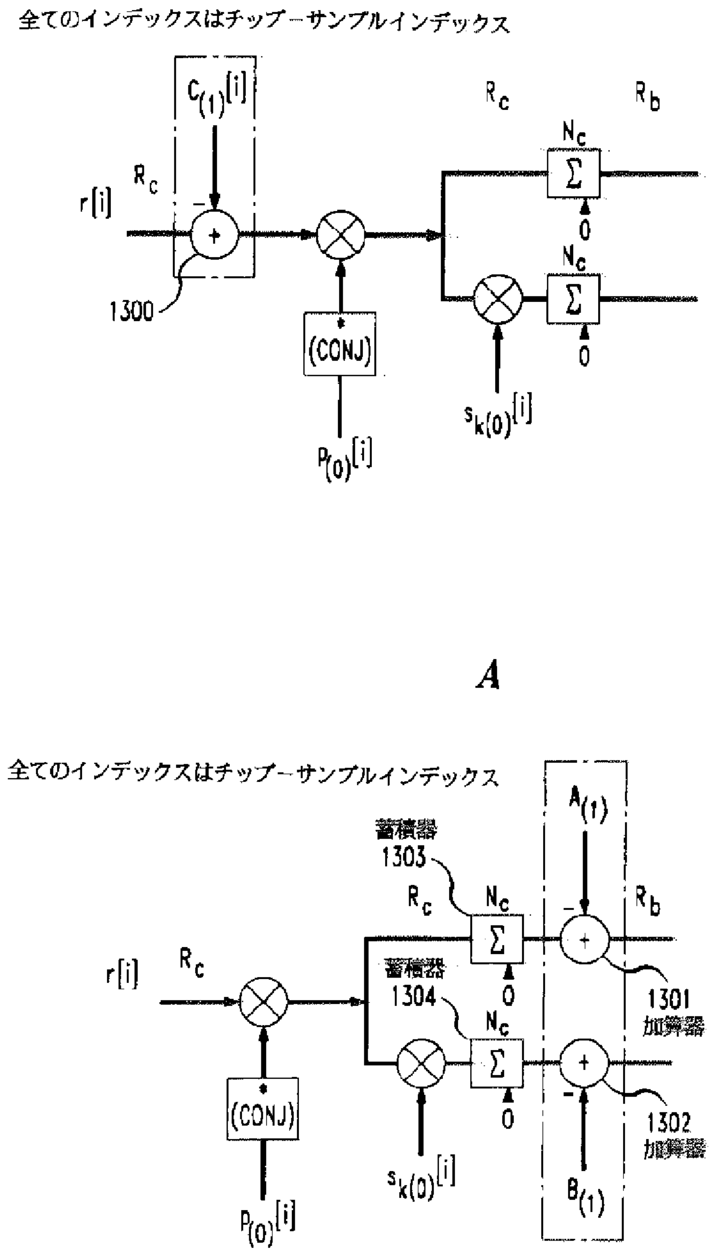
シンボル n-1
シンボル n
シンボル n+1
パス 0
シンボルnのバッファリング
シンボルnの復調

【図１２】

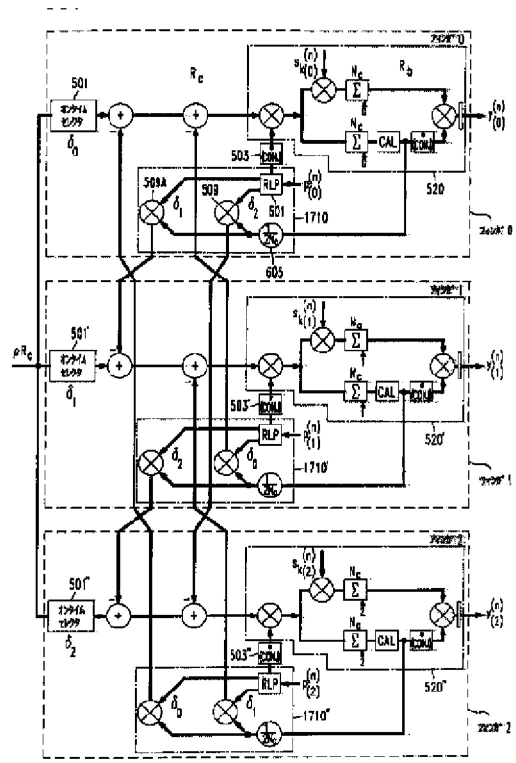


よりよいチャネル予測値を得るためのパイロット除去
除去用チャネル予測
パイロット除去
シンボルバッファ
RLP
CAL
CONJ

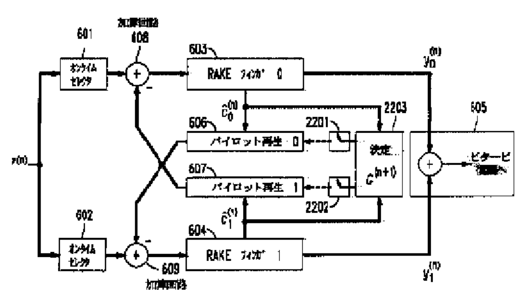
【図１３】



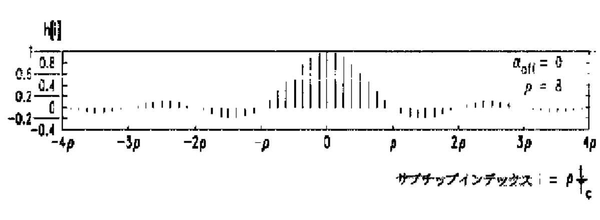
【図１７】



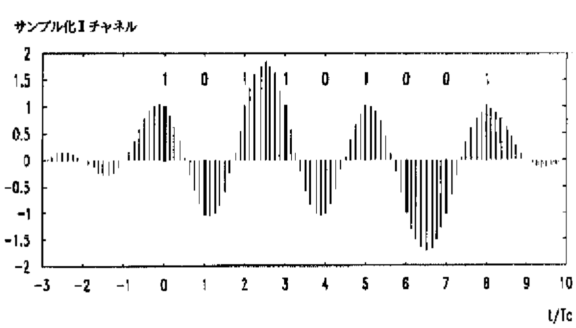
【図２２】



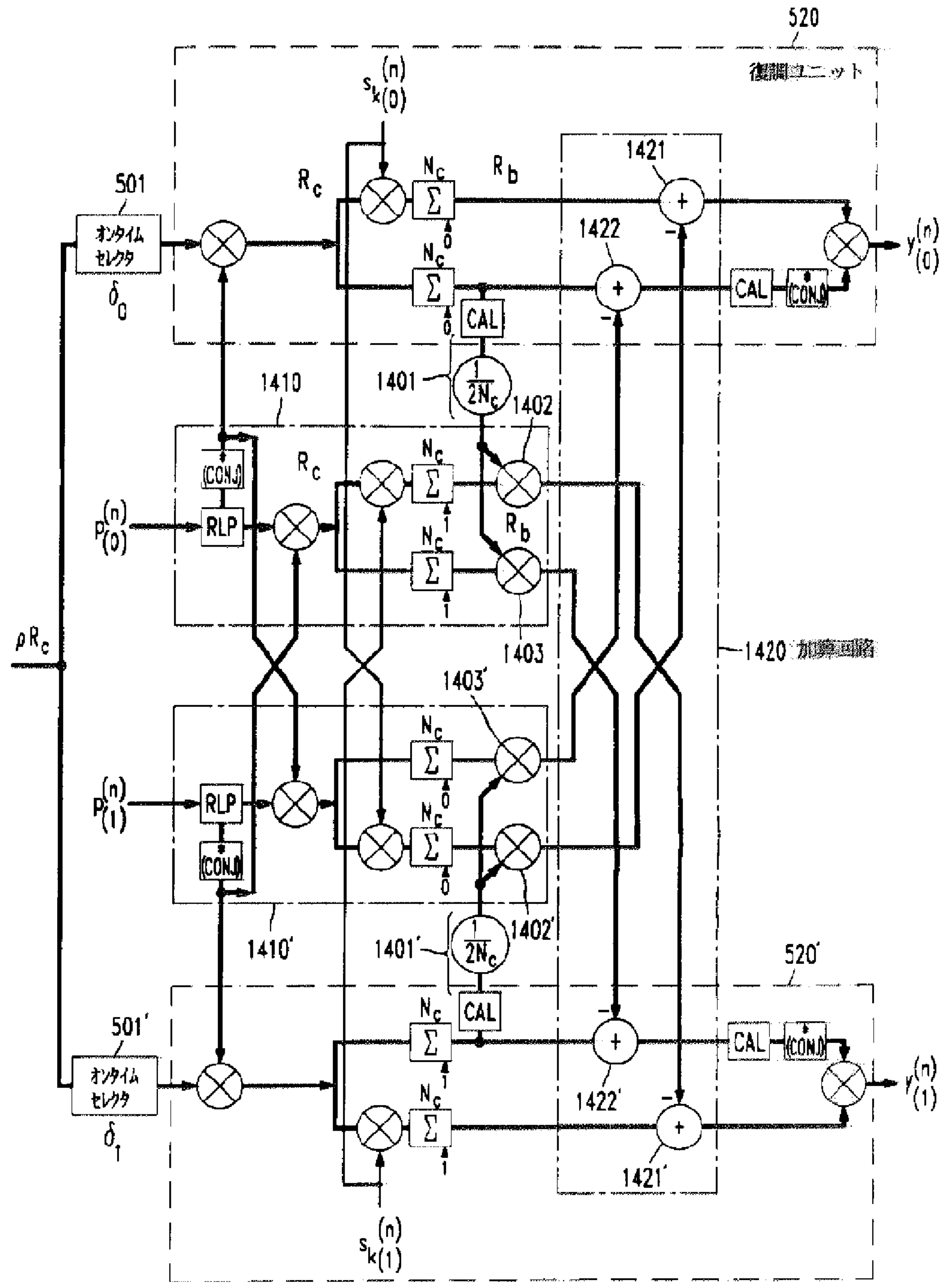
【図１９】



【図２０】

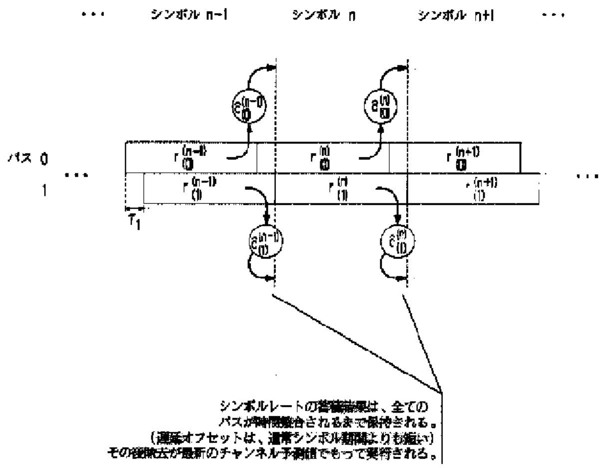


【図１４】

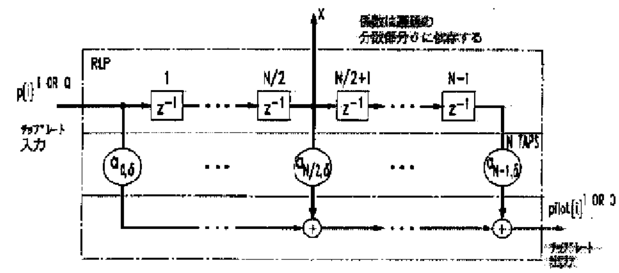


520
復調ユニット
501
オンタイム
セレクタ
δ_0
$s_k^{(n)}(0)$
R_c
N_c
R_b
1421
1422
CAL
(CONJ)
$y^{(n)}(0)$
1410
1401
$\frac{1}{2N_c}$
1402
$p^{(n)}(0)$
RLP
1403
ρR_c
1420 加算回路
1403'
$p^{(n)}(1)$
1410'
1401'
1402'
520'
501'
δ_1
1422'
1421'
$y^{(n)}(1)$
$s_k^{(n)}(1)$

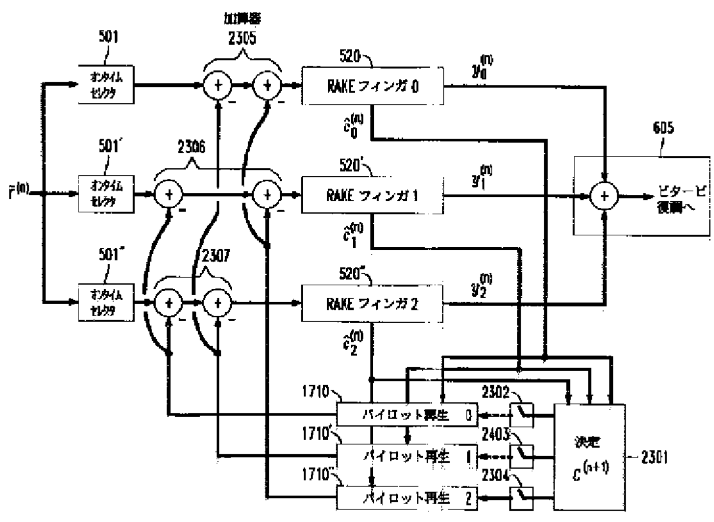
【図１５】


シンボル n−1
シンボル n
シンボル n+1
パス 0
1
シンボルレートの蓄積結果は、全ての
パスが時間整合されるまで保持される。
(遅延オフセットは、通常シンボル期間よりも短い)
その後除去が最新のチャンネル予測値でもって実行される。

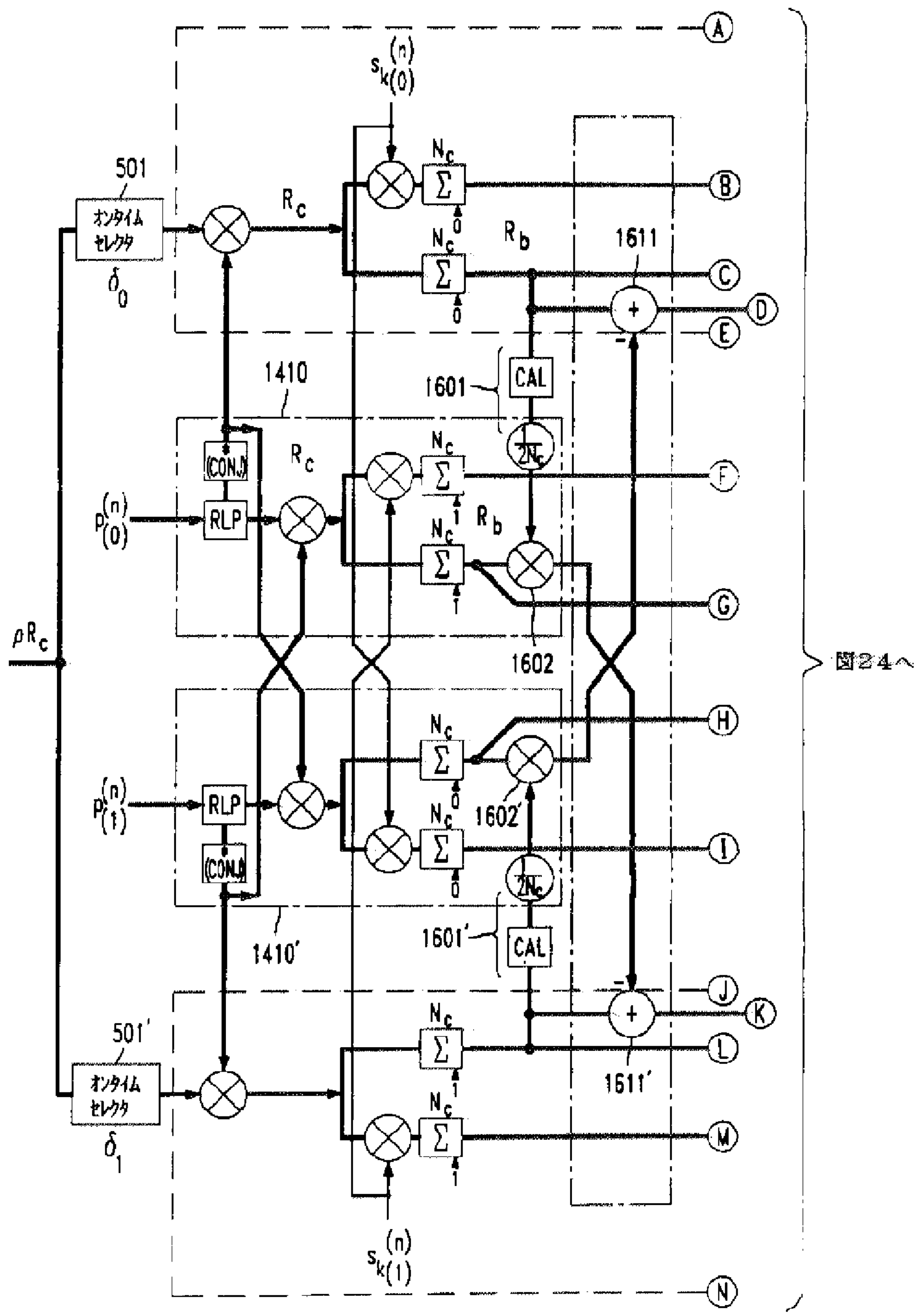
【図２１】


RLP
入力
N TAPS

【図２３】

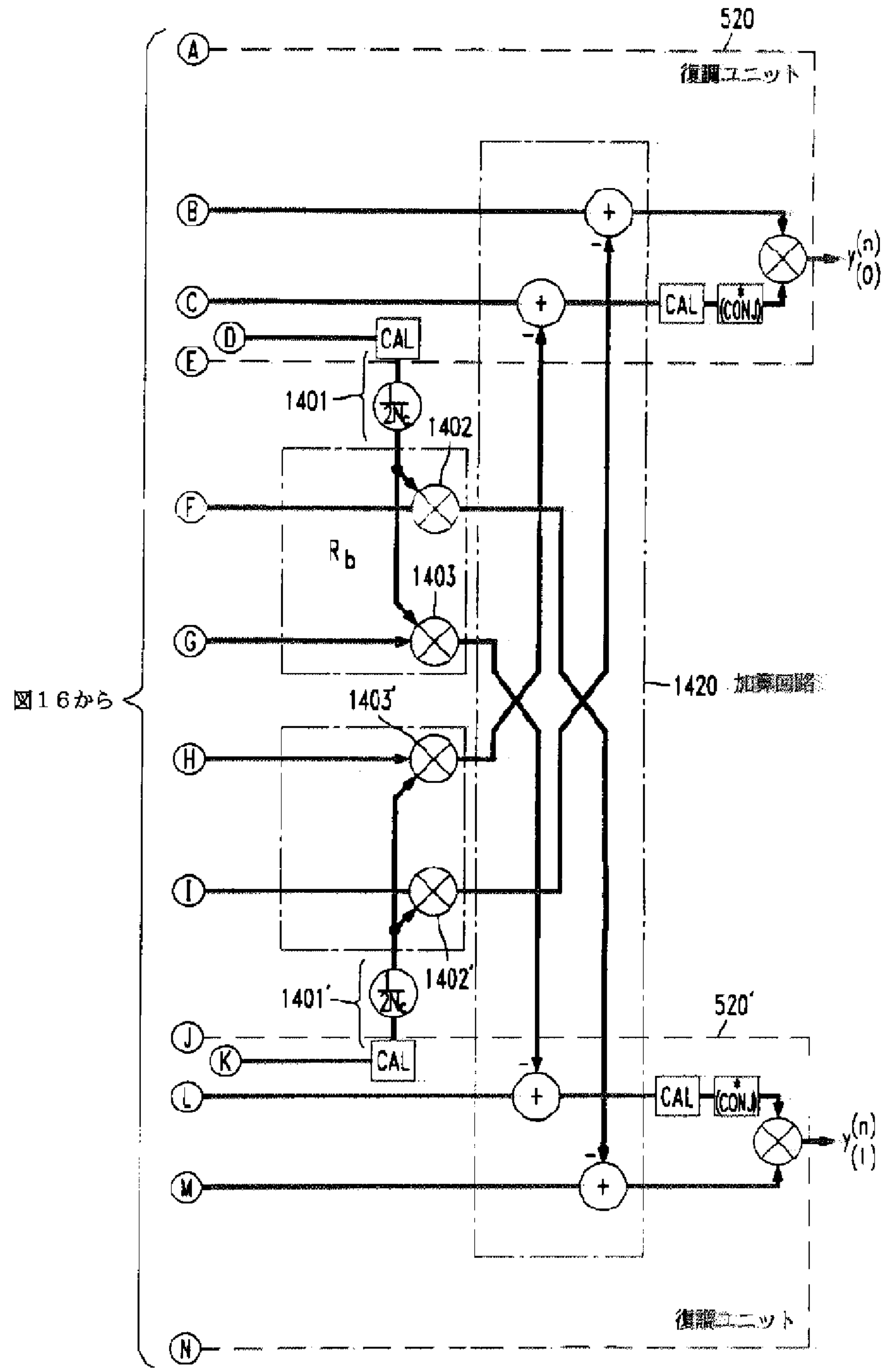

加算器
2305
501
501'
501"
2306
2307
520
520'
520"
RAKE フィンガ 0
RAKE フィンガ 1
RAKE フィンガ 2
605
ビタービ
復調へ
1710
1710'
1710"
パイロット再生 1
パイロット再生 2
2302
2403
2304
決定
2301

【図１６】



501
オンタイム
セレクタ
δ_0
R_c
$s_{k(0)}^{(n)}$
N_c
0
R_b
1611
1410
1601
CAL
$\frac{1}{2N_c}$
(CONJ)
RLP
$p_{(0)}^{(n)}$
1
1602
ρR_c
$p_{(1)}^{(n)}$
1602'
1601'
1410'
501'
δ_1
1611'
$s_{k(1)}^{(n)}$
A
B
C
D
E
F
G
H
I
J
K
L
M
N
図２４へ

【図２４】



フロントページの続き

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JP11252037A

EQUIPMENT AND METHOD FOR ORTHOGONAL FREQUENCY DIVISION MULTIPLEX COMMUNICATION

Publication number : JP11252037A

Date of publication of application : 17.09.1999

Application number : 11-000157

Date of filing : 04.01.1999

Applicant : LUCENT TECHNOL INC

Inventor : D J RICHARD

Abstract:

PROBLEM TO BE SOLVED: To perform operation at a whose back rate such as minimizing the change of a hardware by using the set of a symbol length, quard time and N pieces of sub carriers in a first mode and using the same set of symbol length, quard time and N pieces of sub carriers in a second mode.

SOLUTION: An encoder circuit 1 receives a data stream, divides the data stream into the blocks of continuous groups or bits and introduces redundancy for forward error correction encoding. The block of encoded data bits becomes an input to a high-speed inverse Fourier transforming circuit of complex at N points (N is the number of orthogonal frequency division multiplex sub carriers.) While using phase shift keying of four phase, an IFFT is executed on 2N pieces of encoded data bit blocks received from the encoder circuit 1. A control circuit 4 controls a cyclic perfixer 3 and switches the guard time and symbol period as needed.

(19)日本国特許庁（ＪＰ）　(12) **公開特許公報**（A）　(11)特許出願公開番号

特開平11－252037

(43)公開日　平成11年(1999) 9月17日

(51)Int.Cl.^{6}	識別記号	ＦＩ	
Ｈ０４Ｊ　11/00		Ｈ０４Ｊ　11/00	Ｚ

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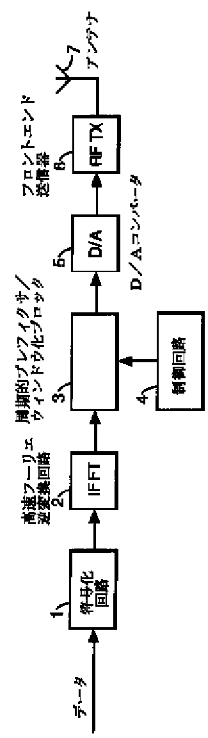
最終頁に続く

(54)【発明の名称】　直交周波数分割多重通信装置とその方法

(57)【要約】

【課題】　送信器または受信器のハードウェアの変更を最小にながらホールバック（代替）レートで動作する装置を提供する。

【解決手段】　本発明によれば、第１シグナリングモード（正常モード）は、シンボル長さＴと、ガードタイムT_Gと、Ｎ個のサブキャリアのセットとを用い、第２モード（ホールバックモード）は、シンボル長さＫＴと、ガードタイムKT_Gと、Ｎ個のサブキャリアの同一のセットを用いる。ここでＫは２以上の整数とする。本発明により、バンド幅とＦＦＴのサイズを変更することなく、ビットレートを下げるだけで、範囲と遅延拡散許容度を増加させることができる。さらにまた、このホールバックレートを用いて、多重アクセスの機能を与えることができる。



【特許請求の範囲】
【請求項１】　時間Ｔの間、直交するサブキャリアの組と、前記サブキャリアの重ね合わせにより表される情報搬送シンボルとを用いる直交周波数分割多重通信装置において、
前記装置は、前記各シンボルの持続時間がＫＴであるような複数のシグナリングモードのうちのひとつのモードで動作し、ここでＫは正整数であり、
前記複数のモードのうち異なるモードは、異なるＫを用いるが、同一のサブキャリアの組を用いることを特徴とする直交周波数分割多重通信装置。
【請求項２】　前記複数のモードのうちの１つのモードは、Ｋ＝１であることを特徴とする請求項１記載の装置。
【請求項３】　ガードタイムが隣接するシンボルの間に挿入され、
前記ガードタイムの長さは、より大きな値のＫのモードよりも大きいことを特徴とする請求項１または２記載の装置。
【請求項４】　前記ガードタイムの長さはＫT_Gであり、前記ＫT_Gは、前記複数のモードの全てに対して同一であることを特徴とする請求項３記載の装置。
【請求項５】　前記装置は受信器であり、前記受信器は前記サブキャリアの重ね合わせから前記シンボルを再生するフーリエ変換手段（１４）と、
Ｋが２以上のモードで動作するときには、持続時間がＴのＫ個の連続する期間の間の平均を取る平均化手段（１５）と、を有することを特徴とする請求項１ないし４のいずれかに記載の装置。
【請求項６】　前記平均化手段（１５）は、前記フーリエ変換手段（１４）の上流側に接続され、
この前記平均化手段（１５）は、持続時間ＫＴのサブキャリアの重ね合わせを受信し、平均された重ね合わせを前記フーリエ変換手段（１４）への入力として取り出すことを特徴とする請求項５記載の装置。
【請求項７】　前記装置は送信器であり、前記送信器は、シンボルを表すサブキャリアの重ね合わせを受信し前記重ね合わせのＫ倍の繰り返しを取り出すことを特徴とする請求項１ないし４のいずれかに記載の装置。
【請求項８】　時間Ｔの間、直交するサブキャリアの組と、前記サブキャリアの重ね合わせにより表される情報搬送シンボルとを用いる直交周波数分割多重通信方法において、
前記各シンボルの持続時間がＫＴであるような所定の複数のシグナリングモードのうちのひとつのモードを選択するステップを含むことを特徴とする直交周波数分割多重通信方法。
【請求項９】　前記複数のモードのうちの１つのモードは、Ｋ＝１であることを特徴とする請求項８記載の方法。
【請求項１０】　ガードタイムが隣接するシンボルの間に挿入され、
前記ガードタイムの長さは、より大きな値のＫのモードよりも大きいことを特徴とする請求項８または９記載の方法。
【請求項１１】　前記ガードタイムの長さはＫT_Gであり、ここでＫT_Gは前記複数のモードの全てに対して同一であることを特徴とする請求項１０記載の方法。
【発明の詳細な説明】
【０００１】
【発明の属する技術分野】本発明は、通信システムに関し、特に直交周波数分割多重化（Orthogonal Frequency Division Multiplexing：ＯＦＤＭ）変調系に関する。
【０００２】
【従来の技術】ＯＦＤＭは、Ｎ個のデータシンボルを１／Ｔの周波数間隔により分離されたＮ個の直交サブキャリアにマッピングするブロック指向の変調系である。ここでＴは、シンボルの持続時間すなわちサブキャリアが直交している時間を意味する。マルチキャリア伝送システムは、ＯＦＤＭ変調を用いて複数のサブキャリア（トーンまたはビンとも称する）を介して、並列に複数のデータビットを送信する。このマルチキャリア伝送の重要な利点は、伝送チャネルにおける信号分散（すなわち遅延拡散）に起因するシンボル間干渉は、後続のシンボルを伝送する間にガードタイム間隔T_Gを挿入することにより減少あるいは除去することができる点である。このためシングルキャリアシステムで必要とされるようなイコライザ（等化器）を取り除くことができる。これは、ＯＦＤＭは、シングルキャリア変調系に対し大きな利点である。ガードタイムにより、意図した信号の後に受信器に到達する各シンボルの遅延コピーは、後続のシンボルを受信する前に消去することができる。このようなＯＦＤＭの魅力的な点については、等化することなし（等化器なしに）にマルチチャネル伝送の欠点を克服できることである。
【０００３】シンボル・ブロックとベースバンドのキャリア信号との間の変換は、通常高速フーリエ変換（ＦＦＴ）技術を用いて実行することができる。このＯＦＤＭに関する議論は Alard and Lasalle 著の EBU Technical Review, no.224, August 1987,Pages 168-190 を参照のこと。様々な通信環境に対してＯＦＤＭの利点を与えるようなフレキシブルなＯＦＤＭシステムが必要とされている。米国特許出願０８／８３４６８４においては、ＯＦＤＭを用いてデータレートを変換（スケーリング）するいくつかの技術を開示している。このスケーリング方法は、クロックレートと、ＦＦＴのサイズと、符号化レートと、信号点配置サイズと、ガードタイムを変更することが関与している。
【０００４】
【発明が解決しようとする課題】本発明の目的は、ハー

ドウェアの変更を最小にするような、ホールバック（代替）レートで動作する装置を提供する。

【０００５】

【課題を解決するための手段】本発明の一実施例によれば、第１シグナリングモード（正常モード）は、シンボル長さＴと、ガードタイムT_Gと、Ｎ個のサブキャリアのセットとを用い、第２モード（ホールバックモード）は、シンボル長さＫＴと、ガードタイムKT_Gと、Ｎ個のサブキャリアの同一のセットを用いる。ここでＫは２以上の整数とする。

【０００６】本発明により、バンド幅とＦＦＴのサイズを変更することなく、ビットレートを下げるだけで、範囲と遅延拡散許容度を増加させせることができる。さらにまた、このホールバックレートを用いて、多重アクセスの機能を与えることができるが、ホールバックレートを用いることは必ずしもスペクトラル効率を悪化させることにはならない。

【０００７】

【発明の実施の形態】図１は、シンボル期間ＴとガードタイムT$_G$をもって伝送されたＯＦＤＭシンボルを表す。ガードタイムT_Gの目的は、分散あるいはマルチパス干渉（総称して以下「遅延拡散」と称する）に起因する連続するシンボル間の干渉を吸収すること、および、このような干渉を受けずにシンボルを受信するためのシンボル持続期間Ｔを残すためである。ある条件下ではあるいはある種のアプリケーションにおいては、ガードタイムT_Gは遅延拡散を吸収するには不十分である場合がある（図１を参照のこと）。より長い期間が必要なこと、すなわち再生された信号内でより高いＳＮ比が必要とされることがある。

【０００８】ガードタイムT_Gを増加させせることは、範囲には影響を及ぼさないが、より長い遅延拡散を吸収できる。クロックレートを減らすことは、ガードタイムT_Gとシンボル期間Ｔを増加させる１つの方法ではあるが、しかしサブキャリア間の周波数間隔１／Ｔを減少させてしまう。このことはチャネルの全体のバンド幅をそれに比例して減少させる。そのため、エアリアス信号を除去するのに必要なフィルタを適応型にしなければならず、そのためハードウェアを変更する必要があることを意味している。

【０００９】図２は、２倍のシンボル期間２Ｔと２倍のガードタイム$2T_G$でもって伝送されるシンボルを示す。この場合、ガードタイムは２倍であり、図に示したシンボル間干渉を吸収できる。シンボル期間が２倍になっているためにＳＮ比および範囲は改善される。しかし、サブキャリアの周波数は、半分にすることができず、またこのことはクロック・レートについても当てはまる。サブキャリアの同一の組は１／Ｔだけ分離（１／２Ｔで分離せず）して用いられる。そのため、チャネルの全体のバンド幅はサブキャリアの周波数の拡散により主に決定され、個々のサブキャリアの幅によりきわめて小さな量に維持されるため、実質的に変化しない。

【００１０】ＯＦＤＭシンボルに対しては、信号はＴ秒（ここでＴはＦＦＴの間隔）後に繰り返すので、受信したシンボルの２つの異なる部分に対し各Ｔ秒の長さにおいて２回のＦＦＴ処理をすることが可能となる。２つのＦＦＴの出力は同一のデータを搬送しているが、異なるノイズを持っているためにそれらは、ＳＮ比が３ｄＢ増加することになる。ＦＦＴは線形操作であるために、Ｔ秒間隔の間まず平均し、この平均された信号を１個のＦＦＴへの入力に用いることができる。このスキームは容易に他のデータレートに拡張することができる。一般的に最高のビットレート以下のＫ倍であるレートは、シンボル期間をＫ倍拡張することにより生成される。シンボルごとにＫ回のＦＦＴ処理を行うことにより、Ｋの処理ゲインが得られ、範囲が増加する。

【００１１】同時に１秒あたりの操作の観点から処理量はホールバックレートに対しては減少する。その理由は平均化された処理はＦＦＴよりもはるかに少ない処理となるからである。例えば６４点のＦＦＴで２μｓのシンボル期間のＯＦＤＭモデムの場合を考えてみる。６４点のＦＦＴは、約１９２個の複素乗算と加算を必要とし、その結果処理ロードは９６Ｍ（百万）演算である。１回の演算は１回の複素乗算とプラス１回の加算として定義される。シンボル期間が２倍になり、ホールバックレートを増加させると、４μｓにおいて６４回の加算と６４点のＦＦＴが行われる。このため処理ロードは、

（１９２＋６４）／４μｓ＝６４Ｍ演算

となる。実際にはこの数字は悪く見積もりすぎている。その理由は余分の加算が乗算として同一の重みを与え、一方それらはハードウェアで行われる場合には、あまり複雑ではないからである。加算は、受信器のほんの一部であり、フルクロックレートで行われる。ＦＦＴとＦＦＴの後の全ての処理（チャネル予測と復号化）はもとのレートよりもＫ倍遅いレートで行われ、これにより電力消費を低減する。

【００１２】図３は、データビットのストリームを受信するＯＦＤＭ送信器を示す。符号化回路１はデータストリームを受信し、それを連続するグループまたはビットのブロックに分ける。符号化回路１は順方向エラー修正符号化用の冗長性を導入する。

【００１３】符号化データビットのブロックは、Ｎ点の複素数の高速フーリエ逆変換回路２への入力である。ここでＮはＯＦＤＭサブキャリアの数である。この実施例においては、４相の位相シフトキーイング（quaternary phase-shift keying：ＱＰＳＫ）を用いてＩＦＦＴが符号化回路１から受信した２Ｎ個の符号化データビットのブロック上で実行される。実際には送信器は送信器の後続のローパスフィルタ処理に起因して、不要な周波数ひずみ（意図するか否かに関わらず）を導入するための

エイリアシングなしにスペクトラムを生成するためにオーバーサンプリングを用いなければならない。オーバーサンプリングを行うために、Ｎ点のＩＦＦＴの代わりにM点のＩＦＦＴが実際に行われる。ただし、Ｍ＞Ｎである。これらの２Ｎ個のビットがＮ個の複素数に変換され、Ｍ－Ｎの入力値は０に設定されたままである。

【００１４】シンボル間干渉に対する感受性を低減するために、周期的プレフィクサ／ウィンドウ化ブロック３がＯＦＤＭシンボルの最後の部分をコピーして、それをＯＦＤＭシンボルのコピーされた部分にプレフィックスすることによりＯＦＤＭシンボルを増加（augment）する。これはサイクリックプレフィクシングと称する。制御回路４は周期的プレフィクサ／ウィンドウ化ブロック３を制御して、必要によりガードタイムとシンボル期間を適宜それらの通常の値T_GとＴの値の間で切り替え、且つホールバック値をKT_GとＫＴの値で切り替える。このホールバック値を与えるためにサイクリックプレフィクサはＯＦＤＭシンボルをＫ－１のコピーでもって増加（augment）しプレフィクスに加える。これは通常のプレフィックスの長さのK倍である。

【００１５】スペクトラムサイドローブを低減するために、周期的プレフィクサ／ウィンドウ化ブロック３は徐々にロールオフするパターンをＯＦＤＭシンボルの振幅に加えることによりＯＦＤＭシンボルに対し、ウィンドイングを実行する。このＯＦＤＭシンボルはＡ／Ｄコンバータに入力され、その後フロントエンド送信器６に送信され、このフロントエンド送信器６がベースバンド波形を適宜ＲＦキャリア周波数に変換してアンテナ７から送信する。

【００１６】図４において、送信されたＯＦＤＭ信号はアンテナ１０を介してＯＦＤＭ受信器により受信される。このＯＦＤＭ信号は、受信回路１１を用いて処理（ダウンコンバート）される。この処理されたＯＦＤＭ信号はＡ／Ｄコンバータ１２に入力される。このデジタルＯＦＤＭ信号をシンボルタイミング回路１３が受信し、このシンボルタイミング回路１３がＯＦＤＭシンボルタイミングを得てタイミング信号を高速フーリエ変換回路１４と積分回路／ダンプフィルタ１５に与える。積分回路／ダンプフィルタ１５はＴ秒だけ分離されたＫ個のサンプルを加える。フィルタのメモリ（M個のサンプルの遅延ラインからなる（ここでMはＦＦＴのサイズである））は、各新たなシンボルの開始時にクリアされる。このリセット時間はシンボルタイミング回路１３によって示され、それはすでに通常のＯＦＤＭ受信器内に入力され、ＦＦＴ間隔のスタートを表す。制御回路１６は平均間隔の数Ｋを設定する。

【００１７】別の実施例においては、積分回路／ダンプフィルタ１５は高速フーリエ変換回路１４の前ではなく後ろに置くこともできる。この場合、各シンボルに対してはK個の連続するＦＦＴの出力が平均化される。しかし、処理負荷は増加する。その理由は、ＦＦＴは常に最大のクロックレートで動かなければならないからである。

【００１８】高速フーリエ変換回路１４により生成されたシンボルのシーケンスは、従来の復号化回路１７に入力されデータ出力信号を生成する。

【００１９】ホールバックレートがもとのレートよりもＫ倍遅いレートで用いられる場合には上記の方法はもとのバンド幅よりもＫ倍小さいバンド幅を有するサブキャリアを生成する。かくして全部の信号のバンド幅は変化しないが、各サブキャリアのバンド幅は小さくなる。これにより同一のバンドで最大Ｋ人のユーザまで周波数分割多重アクセスをすることが可能となる。各ユーザはそのキャリア周波数を１／ＫＴの異なる倍数だけシフトして、他のユーザとの直交性を維持しなければならない。例として６４個のサブキャリアが１ＭＨｚのサブキャリアのスペースでもって用いられた場合には、ホールバックレートをＫ＝４として用いた場合には同一のチャネルに４人のユーザを受け入れることが可能である。これら全ての４人のユーザは、同一の伝送送受信系を用いるが、そのキャリア周波数は０、２５０、５００、７５０ｋＨｚでそれぞれオフセットし、すなわち一般的にはｎ／ＫＴとして表される量だけオフセットする。ｎの値はＫを方とする値で異なる。

【００２０】前掲の特許出願においては制御回路４、１６は外部設定および／または信号の品質をモニタする結果に応じている。同時に前掲の特許出願に議論されているように通信システムのアップリンクとダウンリンクで異なるモードを用いることも可能である。

【００２１】なお、特許請求の範囲に記載した参照番号は発明の容易なる理解のためで、発明を限定的に解釈すべきものではない。

【図面の簡単な説明】

【図１】Ｋ＝１のモードのＯＦＤＭシンボルの伝送状態を表す図。

【図２】本発明によるＫ＝２のモードのＯＦＤＭシンボルの伝送状態を表す図。

【図３】本発明による送信器を表すブロック図。

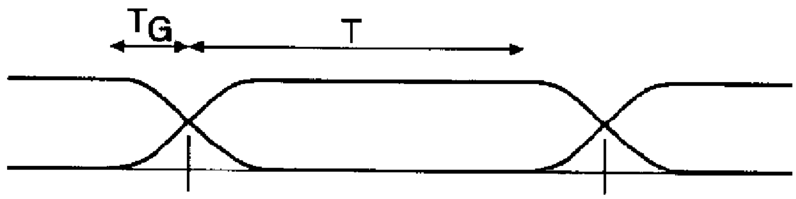
【図４】本発明による受信器を表すブロック図。

【符号の説明】

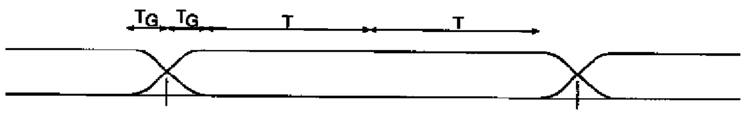
１　符号化回路
２　高速フーリエ逆変換回路
３　周期的プレフィクサ／ウィンドウ化ブロック
４、１６　制御回路
５　Ｄ／Ａコンバータ
６　フロントエンド送信器
７、１０　アンテナ
１１　受信回路
１２　Ａ／Ｄコンバータ
１３　シンボルタイミング回路

１４　高速フーリエ変換回路
１５　積分回路／ダンプフィルタ
１７　復号化回路

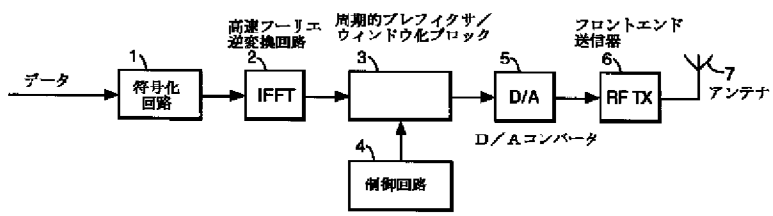
【図１】



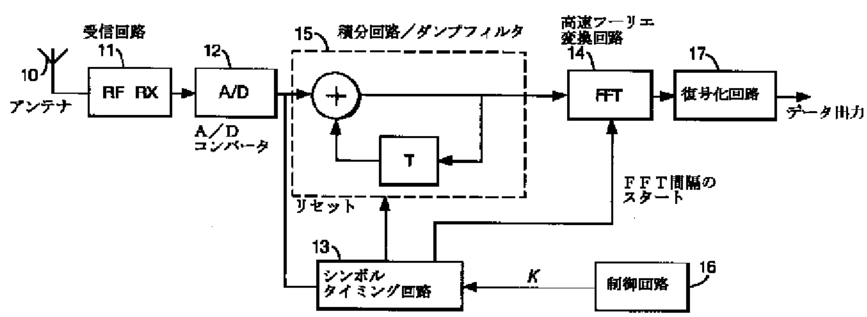
【図２】



【図３】



【図４】



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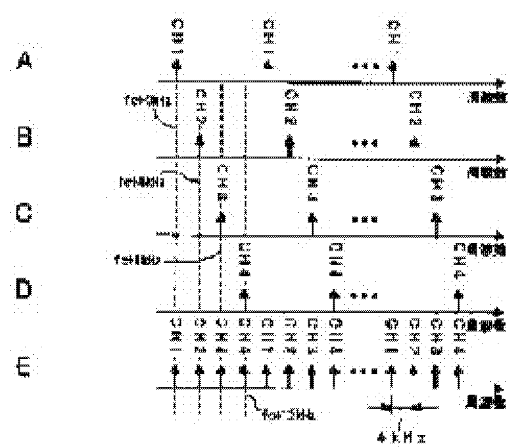
(54) COMMUNICATION METHOD, TRANSMITTER AND RECEIVER

(57)Abstract:

PROBLEM TO BE SOLVED: To allow each receiver or the like to apply communication processing to information with a required minimum processing quantity of itself in the case of multiplexing channels for communication through various transmission routes by each by adopting special arrangement for transmission symbols for each channel on a frequency axis.



SOLUTION: Subcarriers are allocated on a frequency axis for each channel as shown in the following: the subcarriers for a channel 1 are allocated with spacing of 16 kHz from a reference frequency fc (shown in Fig. A), the subcarriers for a channel 2 are allocated with spacing of 16 kHz from a frequency shifted by 4 kHz from the reference frequency fc (shown in Fig. B), the subcarriers for a channel 3 are allocated with spacing of 16 kHz from a frequency shifted by 8 kHz from the reference frequency fc (shown in Fig. C), and the subcarriers for a channel 4 are allocated with spacing of 16 kHz from a frequency shifted by 12 kHz from the reference frequency fc (shown in Fig. D). Signals of each channel are transmitted as a radio wave to

cause the subcarriers to be allocated on a radio transmission channel with spacing of 4 kHz (shown in Fig. E) resulting that signals of the 4 channels are multiplexed and transmitted in one transmission band.

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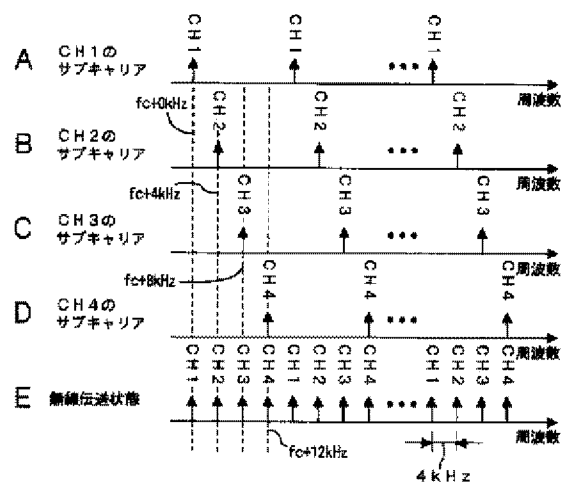
(54)【発明の名称】　通信方法、送信機及び受信機

(57)【要約】
【課題】　様々な伝送レートで通信を行うチャンネルを多重化した際に、各通信は、自らが必要となる必要最低限の処理量をもって、情報の受信などの通信処理を可能とする。
【解決手段】　所定の帯域に複数のチャンネルを設定し、設定したそれぞれのチャンネルでの無線通信を、複数のサブキャリアに送信シンボルを分散させたマルチキャリア信号で行うと共に、各チャンネルでの送信シンボルの周波数軸上での配置を、基準となる周波数間隔に対して２のＮ乗おき（Ｎは正の任意の整数）に配置した。



各チャンネルのサブキャリア配置例

【特許請求の範囲】
【請求項１】　所定の帯域に複数のチャンネルを設定し、
設定したそれぞれのチャンネルでの通信を、複数のサブキャリアに送信シンボルを分散させたマルチキャリア信号で行うと共に、
各チャンネルでの送信シンボルの周波数軸上での配置を、基準となる周波数間隔に対して２のＮ乗おき（Ｎは正の任意の数）に配置した通信方法。
【請求項２】　請求項１記載の通信方法において、
上記通信は無線通信である通信方法。
【請求項３】　請求項１記載の通信方法において、
送信するデータのビットレートに応じて、上記Ｎの値を可変設定した通信方法。
【請求項４】　請求項１記載の通信方法において、
基地局と端末装置との間の通信に適用し、
基地局から送信される下りチャンネルの１チャンネルをパイロットチャンネルとして確保し、残りのチャンネルをトラフィックチャンネルとし、
基地局では、上記パイロットチャンネルで既知信号の送信を行い、
端末装置では、パイロットチャンネルで受信されたシンボルを用いて、上記トラフィックチャンネルで受信したシンボルの伝送路の等化処理を行って、その等化処理されたシンボルの同期検波を行う通信方法。
【請求項５】　請求項１記載の通信方法において、
伝送される信号を、チャンネル単位又は周波数単位で周波数ホッピングさせる通信方法。
【請求項６】　所定の帯域に複数のチャンネルを設定し、
設定したそれぞれのチャンネルでの通信を、複数のサブキャリアに送信シンボルを分散させたマルチキャリア信号で行うと共に、
各チャンネルに割当てられるサブキャリアとして、所定数毎のサブキャリアを使用し、
各チャンネルに割当てられているサブキャリアの隣り合うものどうしで差動変調を行った後に送信し、
受信側では、隣り合うものどうしで差動復調を行う通信方法。
【請求項７】　請求項６記載の通信方法において、
送信側で、各チャンネルに割当てられているサブキャリアの隣り合うものどうしで差動変調を行う代わりに、周波数軸上で隣り合うサブキャリア間で差動変調を行い、
受信側で、各チャンネルに割当てられているサブキャリアの隣り合うものどうしで差動復調を行う代わりに、周波数軸上で隣り合うサブキャリア間で差動復調を行う通信方法。
【請求項８】　複数のサブキャリアに送信シンボルを分散させたマルチキャリア信号を生成させると共に、
上記マルチキャリア信号の１チャンネル内での送信シンボルの周波数軸上での配置を、基準となる周波数間隔に対して２のＮ乗おき（Ｎは正の任意の数）とし、
生成されたマルチキャリア信号を所定の帯域内に設定した複数のチャンネルの内の所定のチャンネルとして送信する送信機。
【請求項９】　請求項８記載の送信機において、
送信するデータのビットレートに応じて、上記Ｎの値を可変設定する送信機。
【請求項１０】　請求項８記載の送信機において、
複数のチャンネルの送信シンボルを個別に生成させた後、１シンボル毎に各チャンネルのシンボルを並べて多重シンボル列を生成し、
生成された多重シンボル列に一括してマルチキャリア信号生成処理を行い、
複数のチャンネルを一括して送信処理を行う送信機。
【請求項１１】　請求項８記載の送信機において、
送信シンボルを生成し、生成した送信シンボルを時間軸上での信号として取り出した後に、自局に割当てられたチャンネルに相当する周波数オフセット分を畳込む処理を行う送信機。
【請求項１２】　請求項８記載の送信機において、
送信される複数のチャンネルの内の１つのチャンネルをパイロットチャンネルとして既知信号を送信処理し、残りのチャンネルをトラフィックチャンネルとして送信処理する送信機。
【請求項１３】　請求項８記載の送信機において、
生成されたマルチキャリア信号を、チャンネル単位又は所定周波数帯域単位で周波数ホッピングさせる周波数ホッピング手段を備えた送信機。
【請求項１４】　複数のサブキャリアに送信シンボルが分散されたマルチキャリア信号を受信し、
１チャンネル内で受信した送信シンボルを、基準となる周波数間隔に対して２のＮ乗おき（Ｎは正の任意の数）の周波数間隔で受信処理する受信機。
【請求項１５】　請求項１４記載の受信機において、
受信した信号より通信に用いられた帯域幅で送信されてきた全シンボル群の内、送信側が送信している通信チャンネルのシンボルのみを抽出し、
この抽出したシンボルをチャンネルデコーダに供給してデコードする受信機。
【請求項１６】　請求項１４記載の受信機において、
受信信号の帯域幅により決定されるサンプルレートにより受信信号のサンプリングを行い、
サンプリングされたシンボルを互いに加算もしくは減算することにより、所望の受信チャンネルを選択して、後段に出力するシンボル数を減少させて、受信時の最大ビットレートにより決定される必要最小限のサンプルレートとし、
この必要最小限のサンプルレートのシンボル数の受信データを受信処理する受信機。

【請求項１７】　請求項１６記載の受信機において、
上記受信データを受信処理する受信処理手段は、最大ビットレートにより決定される処理能力を備え、
上記最大ビットレートよりも低いビットレートでの通信を行う際には所望のビットのみを抽出する受信機。
【請求項１８】　請求項１４記載の受信機において、
パイロットチャンネルの受信処理手段と、トラフィックチャンネルの受信処理手段とを備え、
上記パイロットチャンネルの受信処理手段で受信された既知信号のシンボルを用いて、上記トラフィックチャンネルの受信処理手段で、トラフィックチャンネルの受信シンボルの伝送路の等化処理を行う受信機。
【請求項１９】　請求項１４記載の受信機において、
受信した信号を、チャンネル単位又は所定周波数帯域単位で周波数ホッピングさせる周波数ホッピング手段を備えた受信機。

【発明の詳細な説明】

【０００１】

【発明の属する技術分野】本発明は、例えばセルラ方式による無線電話システムなどの無線通信システムに適用して好適なデジタル無線通信における通信方法と、その通信方法を適用した送信機及び受信機に関する。

【０００２】

【従来の技術】従来、無線電話システムなどのように、広い周波数帯域を複数のユーザでシェアして効率良く通信を行う通信方式としては、例えばＤＳ−ＣＤＭＡ（Direct Sequence-Code Division Multiple Access ）方式がある。このＤＳ−ＣＤＭＡ方式では、送信信号系列を符号により拡散（乗算）し、広帯域信号を生成してこれを送信する。また、受信側では、送信側と同一の拡散符号と受信信号を乗算することにより、逆拡散と呼ばれる効果を得て、受信信号の中から所望の信号成分のみを抽出する。

【０００３】図２７は、従来のＤＳ−ＣＤＭＡ方式を適用したセルラ無線通信システムにおける送信構成を示す。入力端子１に得られる情報ビットストリームは、コーディング部２で符号化ならびにインターリーブなどの処理が施された後に、乗算器３に供給されて、端子３に得られるチャンネル割当ての目的のコードが乗算されて拡散される。拡散されたビットストリームは、次段の乗算器４で、端子４ａに得られるロングコードによりランダム化された後、シンボルマッピング部５で送信シンボルへマッピングされる。このマッピング方法は、通信方式により様々の手法がある。

【０００４】シンボルマッピング部５でマッピングされた送信信号は、必要により加算器６で他の系の送信信号と多重化されて、送信処理部７に供給されて、変調などの高周波処理が行われた後、無線伝送を行う周波数帯域に周波数変換されて、アンテナ８から無線伝送される。

【０００５】ここで入力端子１に得られる情報ビットストリームが例えば８kbpsであるとすると、コーディング部２で符号化率１／２で符号化されて、符号化ビットのビットレートが１６kbpsになり、乗算器３で拡散率６４で拡散すると、１０２４kcps（cps はChip Per Second ）のビットストリームになる。情報ビットストリームのビットレートが異なる場合には、乗算器３での拡散率を変化させれば、送信信号のビットレートを一定にすることができる。

【０００６】また、加算器６で加算する他の送信系についても、加算器６に供給される送信信号のビットストリームが一定であれば、各送信系のコーディング部２に供給される情報ビットストリームとして、種々のものを混在させることができる。

【０００７】次に、従来のＤＳ−ＣＤＭＡ方式で送信処理された信号を受信する構成を、図２８を参照して説明する。アンテナ１１で受信した所定の周波数帯域の信号を、受信処理部１２で中間周波信号などに周波数変換し、この周波数変換された受信信号を復調して、ベースバンドのシンボル系列を得る。このシンボル系列の中から、ビット抽出部１３で受信ビットストリームを抽出する。抽出された受信ビットストリームは乗算器１４に供給して、端子１４ａに得られるロングコードの乗算を行ってデスクランブルすると共に、その乗算器１４の乗算出力を乗算器１５に供給して、端子１５ａに得られる逆拡散コードの乗算を行って逆拡散処理を行い、符号化ビットストリームを得る。そして、その符号化ビットストリームをデコード部１６でデコードして、情報ビットストリームを端子１７に得る。

【０００８】上述した８kbpsの情報ビットストリームが、１０２４kcpsのビットストリームとして送信されている場合の信号を、図２８の構成で受信する場合には、乗算器１５で逆拡散率６４で逆拡散されて、８kbpsの情報ビットストリームが得られる。また、端子１５ａに得られる逆拡散コードの逆拡散率を変化させれば、他のビットレートの情報ビットストリームにも対処できる。

【０００９】ここまでの説明では、ＤＳ−ＣＤＭＡ方式で複数のビットレートの情報ビットストリームを混在させて無線伝送させる場合について説明したが、ＴＤＭＡ（Time Division Multiple Access ）方式で無線伝送させる場合にも、複数のビットレートの情報ビットストリームを混在させることが可能である。図２９は、１フレームがスロット１からスロット８までの８タイムスロットで構成される８ＴＤＭＡ構造の場合の１フレーム構造を示した図である。

【００１０】ここで、１スロット当たりの伝送レートが８kbpsである場合のスロット割当てを想定すると、例えば伝送レート８kbpsのユーザＡ，Ｂには、それぞれスロット１，２を割当て、そのスロット１又は２で伝送レート８kbpsの通信を行う。また、伝送レートが１６kbpsのユーザＣには、スロット３とスロット４の２スロットを

割当て、１６kbpsの通信を行う。また、伝送レートが３２kbpsのユーザＤには、スロット５〜スロット８の４スロットを割当て、３２kbpsの通信を行う。このように各ユーザからの伝送要求時の伝送レートなどに応じて、基地局などが１フレーム内のスロットの各ユーザへの割当て数を可変設定することで、ＴＤＭＡ方式で複数のビットレートの情報ビットストリームを混在させて無線伝送させる対処が可能である。

【００１１】また、ＯＦＤＭ（Orthogonal Frequency Division Multiplex ：直交周波数分割多重）方式と称されるマルチキャリア方式で無線伝送を行う場合には、送信構成として、例えば従来図３０に示す構成で行われていた。この構成は、ＤＡＢ（Digital Audio Broadcasting）と称されるデジタルオーディオ放送に適用されている構成で、端子２１に得られる情報ビットストリームは、コーディング部２で符号化などの処理が施された後に、シンボルマッピング部２３で送信シンボルへマッピングされる。そして、送信シンボルを混合回路２４に供給して、他の送信データと多重化される。ここでの多重化は、単純に直列に連結することで、多重化シンボルストリームを生成させる。例えば、１チャンネル当たり６４kspsのシンボルを、１８チャンネル分多重化すると、多重化されたシンボルストリームの伝送レートは６４ksps×１８＝１１５２kspsとなる。

【００１２】この多重化されたシンボルストリームは、周波数変換部２５での周波数インターリーブによりシンボルの並び替えが行われ、各チャンネルのシンボルがばらばらに並ぶことになる。この並び替えられたシンボルストリームは、逆フーリエ変換回路（ＩＦＦＴ回路）２６で逆フーリエ変換処理により周波数軸上に配置されたマルチキャリア信号となり、このＩＦＦＴ回路２６の出力が送信処理部２７で無線送信処理されて、所定の周波数帯域で無線送信される。

【００１３】このマルチキャリア信号を受信する側の構成としては、図３１に示すように、アンテナ３１で受信した所望の周波数帯域の信号を、受信処理部３２でベースバンド信号とする。ここで、マルチキャリア信号のベースバンド信号成分は、情報が周波数軸上に並んだ信号であるので、高速フーリエ変換回路（ＦＦＴ回路）３２に供給して、フーリエ変換処理を行い、周波数軸上に並んだサブキャリアを抽出する。このとき、フーリエ変換処理によって出力されるシンボルは、受信した信号帯域全体のサブキャリア群となる。

【００１４】このサブキャリア群の変換信号は、シンボル選択部３４に供給して、送信側で行われた周波数インターリーブにより配置された所望のチャンネルのシンボルの存在位置からシンボルを抽出する。さらに、この抽出されたシンボルストリームは、ビット抽出部３５に供給して、符号化ビットストリームを抽出し、この符号化ビットストリームをデコード部３６に供給して、情報ビットストリームを出力端子３７に得る。

【００１５】この従来のＯＦＤＭ方式においては、サブキャリア毎に異なるチャンネルのシンボルを割当てることにより多重化が行われている。従って、受信機が備えるフーリエ変換回路（ＦＦＴ回路）は、多重化されて伝送される全チャンネル分のシンボルを変換処理して、その変換後にチャンネルの選定を行っている。

【００１６】

【発明が解決しようとする課題】上述したＤＳ−ＣＤＭＡ方式を適用したセルラ方式の通信システムでは、使用周波数帯域を固定して、拡散率を可変することにより、可変レートのデータ伝送を可能としている。使用周波数帯域を固定することにより、単一の高周波回路のみで可変ビットレートサービスを提供する端末装置を構成することが可能になっている。

【００１７】しかしながらＤＳ−ＣＤＭＡ方式は、通信制御方式が非常に複雑であり、例えばセルラ方式に適用した場合には、基地局を切換えるハンドオフ処理や、システム内の他の通信との干渉を防止するための送信パワーコントロールなどを、非常に精度良く行う必要がある。また、ＤＳ−ＣＤＭＡ方式は、基本的に全チャンネルが同一の周波数帯域をシェアしており、かつ各チャンネルの直交性がないことから、送信パワーコントロールが正しく行われない端末装置が１台でも存在したとき、システム全体が機能しなくなると言う危険性を有しており、伝送レート可変などの複雑な処理を行うのに適したシステムとは言えない。

【００１８】さらにＤＳ−ＣＤＭＡ方式で伝送レート可変処理を適用した場合には、復調部分に関しては、数kbps程度の低速の伝送レートで通信を行う端末装置であっても、システムで伝送可能な最も高い伝送レートの通信を行う端末装置と同等の演算処理が必要であり、端末装置における演算処理量を大幅に増加させてしまう。

【００１９】一方、上述したＴＤＭＡ方式を適用した通信システムで可変伝送レートを実現する場合、１チャンネル当たりの最大の伝送レートは、基本的には、〔１スロット割当て時のビットレート〕×〔ＴＤＭＡ数〕に限られており、伝送レートの上限と下限はＴＤＭＡ数によって決定されることになる。従って、伝送レートが変化する範囲が、例えば数kbps程度から百kbps程度などのように、非常に大きい場合には、スロット割当てだけでユーザが所望する伝送レートに対応することが事実上不可能である。１フレーム内のタイムスロット数を非常に多くすれば不可能ではないが、通信制御などの点から現実的ではない。

【００２０】また、上述した従来のＯＦＤＭ方式を適用した通信システムで可変伝送レートによる多重化を実現する場合には、サブキャリア毎に異なるチャンネルのシンボルを割当てることにより多重化が行われているため、受信機が備えるフーリエ変換回路は、多重化されて

伝送される全チャンネル分のシンボルを変換処理する必要があり、非常に多くの変換処理が必要である問題があった。

【0021】本発明の目的は、各々が様々な伝送レートで通信を行うチャンネルを多重化した際に、各受信機などでは、自らが必要となる必要最低限の処理量をもって、情報の通信処理を可能とするものである。

【0022】

【課題を解決するための手段】第1の発明の通信方法は、所定の帯域に複数のチャンネルを設定し、設定したそれぞれのチャンネルでの通信を、複数のサブキャリアに送信シンボルを分散させたマルチキャリア信号で行うと共に、各チャンネルでの送信シンボルの周波数軸上での配置を、基準となる周波数間隔に対して２のＮ乗おき（Ｎは正の任意の数）に配置したものである。

【0023】この通信方法によると、各チャンネルが多重化されてマルチキャリア信号となった送信信号には、各チャンネルの送信シンボルが所定の周波数間隔で配置される。

【0024】第2の発明の通信方法は、所定の帯域に複数のチャンネルを設定し、設定したそれぞれのチャンネルでの無線通信を、複数のサブキャリアに送信シンボルを分散させたマルチキャリア信号で行うと共に、各チャンネルに割当てられるサブキャリアとして、所定数毎のサブキャリアを使用し、各チャンネルに割当てられているサブキャリアの隣り合うものどうしで差動変調を行った後に送信し、受信側では、隣り合うものどうしで差動復調を行うようにしたものである。

【0025】この通信方法によると、チャンネル配置としては、所定数毎のサブキャリアを使用したマルチキャリア信号になると共に、各チャンネル毎のサブキャリアの隣り合うものどうしで差動変調が行われることで、各チャンネルの信号だけで送信処理や受信処理が可能になる。

【0026】また本発明の送信機は、複数のサブキャリアに送信シンボルを分散させたマルチキャリア信号を生成させると共に、マルチキャリア信号の１チャンネル内での送信シンボルの周波数軸上での配置を、基準となる周波数間隔に対して２のＮ乗おき（Ｎは正の任意の数）とし、生成されたマルチキャリア信号を所定の帯域内に設定した複数のチャンネルの内の所定のチャンネルとして送信するものである。

【0027】この送信機によると、各チャンネルの送信シンボルが所定の周波数間隔で配置されて、各チャンネルが多重化されたマルチキャリア信号が送信される。

【0028】また本発明の受信機は、複数のサブキャリアに送信シンボルが分散されたマルチキャリア信号を受信し、１チャンネル内で受信した送信シンボルを、基準となる周波数間隔に対して２のＮ乗おき（Ｎは正の任意の数）の周波数間隔で受信処理するものである。

【0029】この受信機によると、各チャンネルの送信シンボルが所定の周波数間隔で配置されて、各チャンネルが多重化されたマルチキャリア信号を受信できる。

【0030】

【発明の実施の形態】以下、本発明の第１の実施の形態を、図１～図４を参照して説明する。

【0031】本実施の形態においては、セルラ方式の無線電話システムに適用した例としてある。図１は、本例のシステムにおける基地局側又は端末装置側の送信構成を示すものである。ここでは、伝送レートとして３２kbps，６４kbps，９６kbps，１２８kbpsの４種類のレートのデータを伝送することができる構成としたものである。

【0032】端子１０１に得られる上述したいずれかの伝送レートの情報ビットストリームは、コーディング部１０２で符号化ならびにインターリーブなどのコーディング処理を行い、符号化率１／２などの所定の符号化率で符号化する。コーディング部１０２で符号化された各ビットは、シンボルマッピング部１０３に供給して、送信シンボルへマッピングする。ここでの送信シンボルへのマッピング処理としては、ＱＰＳＫ処理，８ＰＳＫ処理，１６ＱＡＭ処理などの処理が適用できる。或いは周波数軸上や時間軸上での差動変調が行われる場合もある。

【0033】このシンボルマッピング部１０３で生成された送信シンボルは、ヌルシンボル挿入部１０４に供給する。ヌルシンボル挿入部１０４では、そのときの伝送レートに応じて振幅（エネルギー）が0のシンボルを規則的に挿入して、元の情報ビットストリームの伝送レートに係わらずシンボルレートを最大の伝送レート（ここでは１２８kbpsに対応したレート）に一定とする処理を行う。

【0034】図2は、このヌルシンボルの挿入状態の例を示したもので、○印で示すシンボル位置が、元の伝送データのシンボル位置で、×印で示すシンボル位置が、ヌルシンボル挿入部１０４で挿入した０のシンボルの位置である。例えば情報ビットストリームの伝送レートが３２kbpsの場合には、図２のＡに示すように、元の各シンボル間に、3つのヌルシンボルを挿入して、１２８kbpsに相当するシンボル数（即ち４倍）の伝送データに変換する。また、情報ビットストリームの伝送レートが６４kbpsの場合には、図２のＢに示すように、元の各シンボル間に、１つのヌルシンボルを挿入して、１２８kbpsに相当するシンボル数（即ち2倍）の伝送データに変換する。また、情報ビットストリームの伝送レートが９６kbpsの場合には、図２のＣに示すように、元の３シンボル毎に、１つのヌルシンボルを挿入して、１２８kbpsに相当するシンボル数（即ち4／３倍）の伝送データに変換する。また、情報ビットストリームの伝送レートが１２８kbpsの場合には、図２のＤに示すように、ヌルシン

ボルを挿入せず、そのままのシンボル数の伝送データとする。

【００３５】ここで、ヌルシンボル挿入部１０４でのヌルシンボルの挿入率Rは、次式で定義される。

【００３６】

【数１】挿入率Ｒ＝（Ｍ－Ｄ）／Ｍ

但し、Mはここでの伝送帯域における最大伝送レート（ここでは１２８kbps）であり、Dは該当するチャンネルでの伝送レートである。

【００３７】このヌルシンボル挿入部１０４での処理は、ヌルシンボルの挿入で、シンボルレートが2^N倍（Nは正の任意の数）になるようにコントロールする処理である。但し、図２のCに示す処理、即ち９６kbpsのレートで伝送する場合には、Nの値が整数とはならないが、上述した〔数１〕式に基づいたヌルシンボルの挿入レートＲ＝１／４の規則を用いた処理である。

【００３８】ヌルシンボル挿入部１０４でヌルシンボルが挿入された送信シンボルは、ランダム位相シフト部１０５でランダム位相シフトによるスクランブル処理（或いは他のスクランブル処理）を行い、そのスクランブル処理された送信シンボルを逆フーリエ変換（ＩＦＦＴ）処理部１０６に供給し、逆高速フーリエ変換の演算処理で、時間軸上に配置されたシンボルストリームを、周波数軸上にサブキャリアが配置されたマルチキャリア信号に変換する。逆フーリエ変換処理部１０６で変換された信号は、ガードタイム付加部１０７に供給してガードタイムを付加すると共に、窓がけ処理部１０８で所定単位毎の信号に送信用の窓がけデータを乗算する。窓がけデータが乗算された送信信号は、送信処理部１０９に供給して、高周波信号を畳込み所定の伝送周波数帯域に周波数変換し、その周波数変換された送信信号をアンテナ１１０から無線送信する。

【００３９】このような構成で無線送信される信号を端末装置又は基地局で受信する構成を、図３を示す。アンテナ１１１が接続された受信処理部１１２では、所定の伝送周波数帯域の信号を受信して、ベースバンド信号に変換する。変換されたベースバンド信号は、窓がけ処理部１１３に供給して、所定単位毎の信号に受信用の窓がけデータを乗算した後、フーリエ変換（ＦＦＴ）処理部１１４に供給し、周波数軸上に配置されたサブキャリアを時間軸上に配置されたシンボルストリームに変換する。

【００４０】変換されたシンボルストリームは、デスクランブル部１１５で送信時のスクランブル処理とは逆のデスクランブル処理を行う。このデスクランブルされたシンボルストリームは、シンボル選択部１１６に供給する。シンボル選択部１１６では、送信時にヌルシンボル挿入部１０４（図１参照）で挿入されたヌルシンボル以外のシンボルを選択（即ちヌルシンボルを除去）する処理を行う。このヌルシンボルが除去されたシンボルストリームをビット抽出部１１７に供給し、符号化ビットを抽出し、その抽出されたビットデータをデコード部１１８に供給してデコードし、デコードされた情報ビットストリームを端子１１９に得る。

【００４１】シンボル選択部１１６で抽出するシンボルとしては、伝送される情報ビットストリームの伝送レートにより異なる。即ち、図２に示すように送信時に挿入された振幅が０のヌルシンボルの位置は、伝送レートにより変化し、それぞれの伝送レートの場合に、○印で示したシンボルだけを抽出する処理を行う。この処理を行うことで、３２kbpsから１２８kbpsまでの伝送レートの伝送を、同じ通信帯域幅を使用して行える。

【００４２】ここでは、３２kbpsから１２８kbpsまでの可変伝送レートで伝送する場合について説明したが、同様の処理により、最大ビット数Mkbpsの通信が行える帯域において、M／2^N kbpsの通信を行うことが可能である。この場合、送信側において、生成されたシンボルとヌルシンボルとは、次の表１に示すパターンで挿入される。この表１において、白丸で示すシンボルは、情報ビットにより生成されたシンボルであり、黒丸で示すシンボルは、ヌルシンボルである。

【００４３】

【表１】

M / 2^0 kbps 通信時	○○○○○○○○○○○○○○○○
M / 2^1 kbps 通信時	○●○●○●○●○●○●○●○●
M / 2^2 kbps 通信時	○●●●○●●●○●●●○●●●
M / 2^3 kbps 通信時	○●●●●●●●○●●●●●●●
M / 2^4 kbps 通信時	○●●●●●●●●●●●●●●●

(○：情報ビットより生成されたシンボル，●：ヌルシンボル)

【００４４】以上のような通信を行うことで、低速伝送から高速伝送までを同じ通信帯域幅を用いて行うことが可能となり、例えば単一の高周波回路（送信処理回路や受信処理回路）のみしか備えていない端末装置においても可変伝送レートの通信が可能になる。

【００４５】なお、この第１の実施の形態で説明した伝送処理を、ＴＤＭＡ構造で行うようにすることで、最低伝送レートと最大伝送レートとの差をより大きくすることが可能になる。図４は、この場合のフレーム構造の例を示す図で、例えばスロット１～スロット８の８タイムスロットで１フレームが構成される８ＴＤＭＡで構成されている場合に、１つのスロットで３２kbps（ヌルシン

ボル挿入率R＝３／４）から１２８kbps（ヌルシンボル挿入率R＝０／４）までのレートのマルチキャリア信号の伝送が可能な帯域が設定してあるとすると、１フレームで１スロットだけを使用した通信では、３２kbpsから１２８kbpsのレートでの伝送が行われ、１フレームの２スロットを使用した通信では、２５６kbpsのレートまでの伝送が行われ、以下使用するスロット数を増やすことで、最大で８スロットを使用して、ヌルシンボル挿入率R＝０／４としたとき１２８kbps×８＝１０２４kbpsの伝送レートでの通信が可能となる。

【００４６】また、この第１の実施の形態で説明した伝送処理でヌルシンボルを挿入した箇所（ヌルシンボルによるサブキャリア）は、他の系の通信で使用することができる。このようにヌルシンボルの挿入位置のサブキャリアを、他の通信に使用することで、多重通信を効率良く行うことができる。例えば、図１に示す送信処理で、６４kbpsのレートの情報ビットストリームを送信する際には、ヌルシンボルの挿入位置で、他の系の通信を行うことで、２つの系の６４kbpsのレートの情報ビットストリームの伝送が、１つの伝送帯域で可能である。同様に、３２kbpsのレートの場合には、４つの系の３２kbpsのレートの情報ビットストリームの伝送が、１つの伝送帯域で可能である。さらに、９６kbpsのレートの伝送と、３２kbpsのレートの伝送とを、１つの伝送帯域で行うこともできる。

【００４７】次に、本発明の第２の実施の形態を、図５～図７を参照して説明する。本実施の形態においても、セルラ方式の無線電話システムに適用した例としてあり、この例では１つの送信機から多重送信を行うようにしたものである。この多重送信は、例えば基地局から複数の系の送信信号を同時に送信する場合に適用できる。この実施の形態において、多重通信を行う構成以外は、上述した第１の実施の形態で説明した処理と基本的に同じであり、受信系の構成については省略する。

【００４８】図5は、本実施の形態での送信構成を示した図である。ここでは、チャンネル１，チャンネル２…チャンネルN（Nは任意の整数）のチャンネル数Nの情報ビットストリームが、端子１２１ａ，１２１ｂ…１２１ｎに得られるものとする。各端子１２１ａ～１２１ｎに得られる各チャンネルの情報ビットストリームは、ここでは同じ伝送レートのビットストリームとしてあり、それぞれ別のコーディング部１２２ａ，１２２ｂ…１２２ｎに供給して、符号化ならびにインターリーブなどのコーディング処理を個別に行う。コーディング部１２２ａ～１２２ｎで符号化された各チャンネルのビットストリームは、それぞれ別のシンボルマッピング部１２３ａ，１２３ｂ…１２３ｎに供給して、各チャンネル毎に個別に送信シンボルへマッピングする。ここでの送信シンボルへのマッピング処理としては、ＱＰＳＫ処理，８ＰＳＫ処理，１６ＱＡＭ処理などの処理が適用できる。或いは周波数軸上や時間軸上での差動変調が行われる場合もある。

【００４９】各チャンネル毎のシンボルマッピング部１２３ａ～１２３ｎで生成された送信シンボルは、混合回路（マルチプレクサ）１２４に供給して、１系統のシンボルストリームに混合する。図6は、混合回路１２４での処理の概念を簡単に示す図で、ここでは例えばチャンネル１～チャンネル４のチャンネル数４のシンボルストリームを、１系統のシンボルストリームに変換するものである。チャンネル１のシンボルストリームが混合回路１２４の端子１２４ａに得られ、チャンネル２のシンボルストリームが混合回路１２４の端子１２４ｂに得られ、チャンネル３のシンボルストリームが混合回路１２４の端子１２４ｃに得られ、チャンネル４のシンボルストリームが混合回路１２４の端子１２４ｄに得られる。このとき、混合回路１２４を構成するスイッチの接点１２４ｍが、各端子１２４ａ～１２４ｄを順に周期的に選択する処理を行って出力する。

【００５０】図7は、この混合状態の例を示した図で、例えば図７のＡ，Ｂ，Ｃ，Ｄに示す状態で、それぞれ別のチャンネル１，２，３，４のシンボルストリームが得られるとき、各チャンネルのシンボルを順に選択して、図７のＥに示す１系統の混合ストリームを得る。例えば、各チャンネルのストリームが、３２kbpsのレートの情報ビットストリームのシンボルであるとき、１２８kbpsのレートの情報ビットストリームに相当するシンボルストリームとなる。なお、各チャンネルのシンボルの送出タイミングが同期してない場合には、バッファメモリなどを使用した同期処理が必要になる。

【００５１】図5の説明に戻ると、混合回路１２４で混合された送信シンボルは、ランダム位相シフト部１２５でランダム位相シフトによるスクランブル処理（或いは他のスクランブル処理）を行い、そのスクランブル処理された送信シンボルを逆フーリエ変換（ＩＦＦＴ）処理部１２６に供給し、逆高速フーリエ変換の演算処理で、時間軸上に配置されたシンボルストリームを、周波数軸上にサブキャリアが配置されたマルチキャリア信号に変換する。逆フーリエ変換処理部１２６で変換された信号は、ガードタイム付加部１２７に供給してガードタイムを付加すると共に、窓がけ処理部１２８で所定単位毎の信号に送信用の窓がけデータを乗算する。窓がけデータが乗算された送信信号は、送信処理部１２９に供給して、高周波信号を畳込み所定の伝送周波数帯域に周波数変換し、その周波数変換された送信信号をアンテナ１３０から無線送信する。

【００５２】このように無線送信される信号を受信する側（例えば基地局からの信号を受信する端末装置）では、例えば上述した第１の実施の形態で説明した図3の構成で受信処理を行うことで、任意のチャンネルの信号を抽出して処理できる。

【0053】なお、ここでは4チャンネルの多重化を行う場合を例として説明したため、多重化されたシンボルストリーム（図7のE）での各チャンネルのシンボルの出現周期は4となっているが、最大のチャンネル多重数はこれに限定されるものではない。最大のチャンネル多重数は、2^{n}（ここでのnは正の整数：即ちn＝1，2，3，4・・・）と設定することができ、この場合の各チャンネルのシンボルの出現周期は、最大の多重数と同じ2^{n}となる。実際の通信で使用するチャンネル数が、最大の多重数よりも小さい場合には、使われてないチャンネルのシンボルとして、第1の実施の形態で説明したヌルシンボル（振幅が0のシンボル）を挿入すれば良い。

【0054】次に、本発明の第3の実施の形態を、図8及び図9を参照して説明する。本実施の形態においても、セルラ方式の無線電話システムに適用した例としてあり、この例でも第2の実施の形態と同様に、1つの送信機から多重送信を行うようにしたものであり、第2の実施の形態に対応する部分には同一符号を付し、その詳細説明は省略する。

【0055】ここで本実施の形態の場合には、各チャンネルの伝送レートが異なる場合の例としてあり、図8は本実施の形態での送信構成を示した図である。ここでは、チャンネル1，チャンネル2，チャンネル3のチャンネル数3の情報ビットストリームが、端子131a，131b，131cに得られるものとする。各チャンネルの伝送レートとしては、例えばチャンネル1，チャンネル2がそれぞれ32kbpsであり、チャンネル3が64kbpsであるとする。各端子131a〜131cに得られる各チャンネルの情報ビットストリームは、それぞれ別のコーディング部132a，132b，132cに供給して、符号化ならびにインターリーブなどのコーディング処理を個別に行う。コーディング部132a，132bで符号化されたチャンネル1，チャンネル2のビットストリームは、それぞれのチャンネル用のシンボルマッピング部133a，133bに供給して、各チャンネル毎に個別に送信シンボルへマッピングする。また、チャンネル3のビットストリームは、2つの系統のビットストリームに2分割し、一方の系統のビットストリームはシンボルマッピング部133cに供給すると共に、他方の系統のビットストリームはシンボルマッピング部133dに供給し、それぞれ別に送信シンボルへマッピングする。

【0056】各シンボルマッピング部133a〜133dでマッピングされた送信シンボルは、混合回路134に供給して、1系統に多重化する。図9は、ここでの多重化状態の例を示してあり、2つの系統に分割されたチャンネル3のシンボルストリームを、同じ間隔で周期的に配置すると共に、その間にチャンネル1のシンボルストリームとチャンネル2のシンボルストリームを周期的に配置する。即ち、例えばチャンネル1，チャンネル3，チャンネル2，チャンネル3・・・の配置を繰り返し設定する。

【0057】この多重化されたシンボルストリームは、ランダム位相シフト部125でランダム位相シフトによるスクランブル処理（或いは他のスクランブル処理）を行い、そのスクランブル処理された送信シンボルを逆フーリエ変換（IFFT）処理部126に供給し、逆高速フーリエ変換の演算処理で、時間軸上に配置されたシンボルストリームを、周波数軸上にサブキャリアが配置されたマルチキャリア信号に変換する。逆フーリエ変換処理部126で変換された信号は、ガードタイム付加部127に供給してガードタイムを付加すると共に、窓がけ処理部128で所定単位毎の信号に送信用の窓がけデータを乗算する。窓がけデータが乗算された送信信号は、送信処理部129に供給して、高周波信号を畳込み所定の伝送周波数帯域に周波数変換し、その周波数変換された送信信号をアンテナ130から無線送信する。

【0058】このように無線送信される信号を受信する側（例えば基地局からの信号を受信する端末装置）では、例えば上述した第1の実施の形態で説明した図3の構成で受信処理を行うことで、任意のチャンネルの信号を抽出して処理できる。即ち、図9に示す状態で多重化された伝送信号から、チャンネル1又はチャンネル2の信号を抽出する場合には、4周期毎のシンボルを抽出することで、そのチャンネルの信号が受信でき、チャンネル3の信号を抽出する場合には、2周期毎のシンボルを抽出することで、そのチャンネルの信号が受信できる。

【0059】なお、ここでは最大128kbpsまで伝送できる帯域で、32kbpsと64kbpsの伝送レートを混在させて通信を行う例として説明したが、これに限定されるものではない。即ち、各チャンネルの伝送レートD〔kbps〕は、基本的には次式のように設定できる。

【0060】

【数2】伝送レートD＝M／2^{N}〔kbps〕

ここで、N＝1，2，3・・・の正の整数、Mは該当する帯域における最大伝送レートである。

【0061】また、第1の実施の形態で説明した96kbpsのように、〔数2〕式で設定されるレートの間の値のレートを設定しても良い。

【0062】次に、本発明の第4の実施の形態を、図10〜図15を参照して説明する。本実施の形態においても、セルラ方式の無線電話システムに適用した例としてあり、この例では複数の送信機から多重送信を行うようにしたものである。例えば、複数の端末装置から同時に多重送信を行って、基地局で一括して受信する場合が相当する。

【0063】図10は本実施の形態での送信構成を示した図である。ここでは、チャンネル1〜チャンネルN（Nは任意の整数）の情報ビットストリームが、それぞ

れ別の送信機の端子141a〜141nに個別に得られるものとする。各送信機は基本的には共通の構成であり、チャンネル1の信号を処理する送信機の構成を説明すると、端子141aに得られる情報ビットストリームは、コーディング部142aで符号化ならびにインターリーブなどのコーディング処理を行う。コーディング部142aで符号化された各ビットは、シンボルマッピング部143aに供給して、送信シンボルへマッピングする。

【0064】このシンボルマッピング部143aで生成された送信シンボルは、ランダム位相シフト部144aでランダム位相シフトによるスクランブル処理（或いは他のスクランブル処理）を行い、そのスクランブル処理された送信シンボルを逆フーリエ変換（IFFT）処理部145aに供給し、逆高速フーリエ変換の演算処理で、時間軸上に配置されたシンボルストリームを、周波数軸上にサブキャリアが配置されたマルチキャリア信号に変換する。逆フーリエ変換処理部145aで変換された信号は、内部チャンネル選択部146aで内部チャンネル選択処理が行われ、この内部チャンネル選択処理が行われたマルチキャリア信号を、送信処理部147aに供給して、高周波信号を畳込み所定の伝送周波数帯域に周波数変換し、その周波数変換された送信信号をアンテナ148aから無線送信する。

【0065】内部チャンネル選択部146aの構成を図11に示す。前段の回路から端子151に得られる信号を、シンボル繰り返し部152に供給し、そのときの伝送レートに応じて数のシンボル反復処理を行う。例えば、ここでの1伝送帯域での最大伝送レートが128kbpsで、無線伝送されるマルチキャリア信号の伝送路上でのサブキャリア間隔を4kHz間隔とし、1チャンネルでの伝送レートが32kbpsであるとする。このとき、前段の逆フーリエ変換処理部145aでは、サブキャリア間隔が16kHzのマルチキャリア信号への変換処理を行う。

【0066】シンボル繰り返し部152では、この信号のシンボル成分を4倍に反復する処理を行い、4kHz間隔の信号に変換する。例えば図11に示すように、シンボル繰り返し部152の入力部に示した波形が、このシンボル繰り返し部152で4回反復された波形に変換されている。この逆フーリエ変換されたシンボルストリームを多重分繰り返すことによって、該当するチャンネルが使用していないサブキャリアにヌルシンボルを挿入することと等価の効果を得ることになる。

【0067】このシンボル繰り返し部152で繰り返された信号は、乗算器153で、オフセット周波数発生器154が出力するオフセット周波数と乗算される。この乗算により、該当するチャンネルの周波数オフセット分、各シンボルに位相の旋回が生じることになる。なお、該当するチャンネルの周波数オフセットが0Hzである場合には、定数との乗算になる。即ち、この乗算器153で乗算されたシンボル系列によって、どのチャンネルに割当てられたサブキャリアを使用するかが決定される。オフセット周波数が乗算された信号は、窓がけ処理部155に供給して、所定単位毎に送信用の窓がけデータを乗算し、端子156から送信処理部147aに供給する。

【0068】各チャンネルで送信処理される信号の状態の例を図12に示す。ここでは、1伝送帯域での最大伝送レートが128kbpsで、この128kbpsの伝送レートのデータを、4kHz間隔のサブキャリアによるマルチキャリア信号により伝送される構成としてある場合に、4つの送信機から1つの伝送帯域を使用して、それぞれの送信機から伝送レートが32kbpsのデータを、この1伝送帯域に多重伝送する場合を示したものである。

【0069】図12のA，B，C，Dは、それぞれ各送信機から送信されるチャンネル1，チャンネル2，チャンネル3，チャンネル4の送信信号を示したもので、各チャンネルの信号は、サブキャリアが16kHz間隔のマルチキャリア信号としてある。ここで、各チャンネルでサブキャリアが存在する周波数位置は、チャンネル1が図12のAに示すように、基準となる周波数fcから16kHz間隔としてあり、チャンネル2が図12のBに示すように、周波数fcから4kHzシフトした周波数位置から16kHz間隔としてあり、チャンネル3が図12のCに示すように、周波数fcから8kHzシフトした周波数位置から16kHz間隔としてあり、チャンネル4が図12のDに示すように、周波数fcから12kHzシフトした周波数位置から16kHz間隔としてある。

【0070】これらの各チャンネルの信号が無線送信されることで、無線伝送路上では図12のEに示すように、4kHz間隔でサブキャリアが配置された状態となり、1つの伝送帯域に4つのチャンネルの信号が多重伝送されることになる。この場合、各送信機が備える逆フーリエ変換処理部での高速逆フーリエ変換処理としては、そのチャンネルで扱う32kbpsの伝送レートの信号を16kHz幅のサブキャリア群に変換する処理だけで良く、逆フーリエ変換処理部での処理量を、そのシステムにおけるサブキャリア間隔で必要な処理量よりも大幅に少なくすることができる。

【0071】ここでは、32kbpsの伝送レートの信号の通信を行う例について説明したが、例えば同じ伝送帯域で64kbpsの伝送レートの信号の通信を行う場合には、そのレートの通信に見合う規模の逆フーリエ変換処理部により演算を行い（即ち32kbpsの通信の時に比べて倍のサンプル数が出力される）、内部チャンネル選択部でのシンボル反復で2倍に反復すれば良く、どのような伝送レートの場合でも同様の処理で送信信号の生成が可能である。この場合、各送信機（端末装置）が備える処理